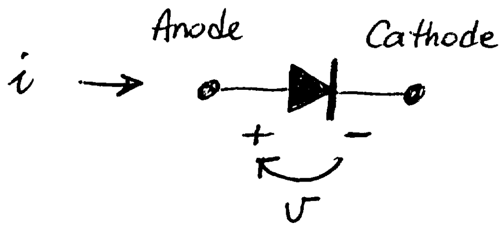

CLASS NOTES

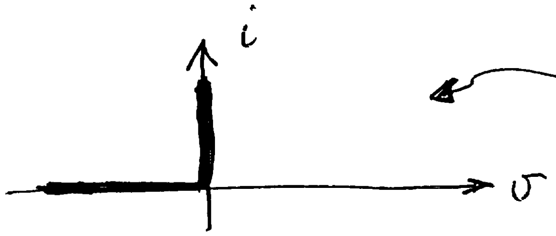
CHAPTER 5

SEMICONDUCTOR DIODES

The diode is a non linear device. It has a non linear i-v characteristic.
It exhibit a behavior which is dependent on the direction of the applied voltage.

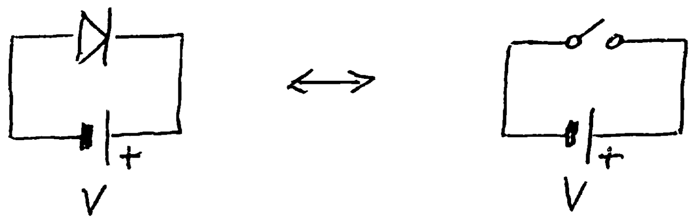


The ideal diode

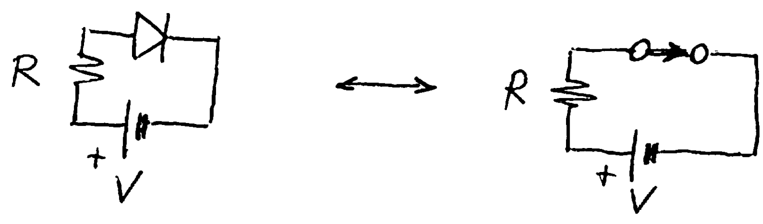


Rapid analysis of circuits using diodes (we fine-tune the analysis later)

if a negative voltage is applied to the diode, no current flows and the diode behaves like an open circuit.
A diode operated in the reverse direction (reverse biased) is said to be <u>CUT OFF</u>.

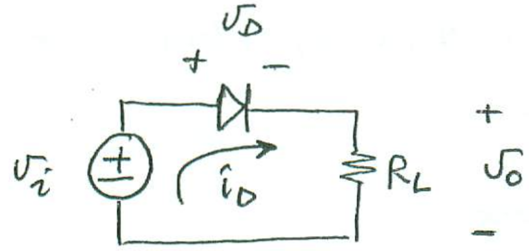


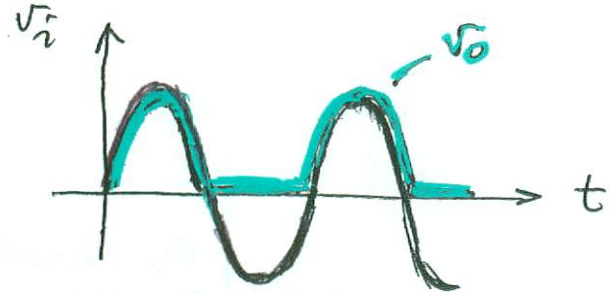
If a positive voltage is applied the ideal diode behave like a short circuit and current flows through it.
A diode eperated in the forward direction (forward biased) is said to be <u>ON</u>.

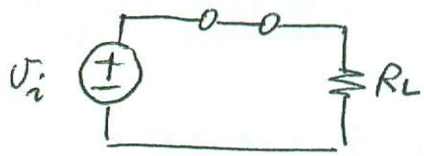


let's see a couple of interesting ways we could use a circuit element with such a behavior:

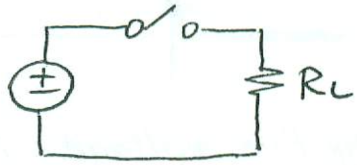
RECTIFIER





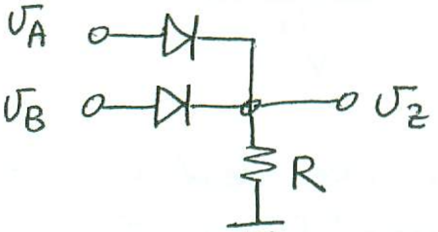


$v_i \geq 0$

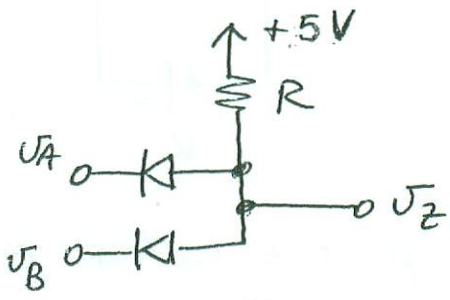


$v_i \leq 0$

DIODE LOGIC GATES:



OR gate



AND gate

Diode current-voltage characteristic

Theoretical analysis of a pn junction results in the following i-v relationship:

$$i = I_S \left(e^{v/\eta V_T} - 1 \right)$$

$V_T = \dfrac{KT}{q}$ = thermal voltage $\left\{ \begin{array}{l} \text{at room temperature} \\ 27^\circ C \ (= 300^\circ K) \text{ the} \\ \text{value of } V_T = 26\,mV \end{array} \right.$

K = Boltzmann's constant $= 1.38 \times 10^{-23}$ joules/kelvin

T = temperature in kelvin = 273 + temperature in °C

q = magnitude of electron charge $= 1.6 \times 10^{-19}$ Coulomb

η = empirical scaling constant = exponential ideality factor

I_S = saturation current

η has a value between $0.5 \div 2$. It depends on the material (type of semiconductor used and the doping).
In general we will assume $\eta = 1$ unless otherwise specified

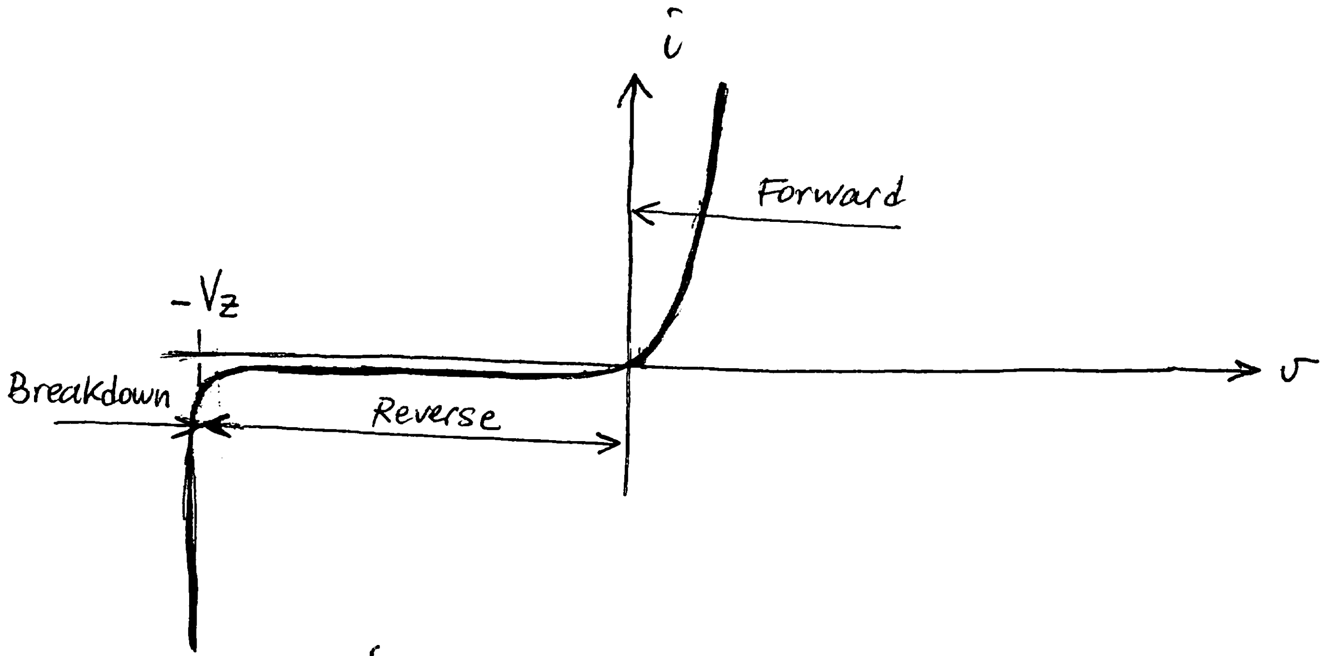
I_S is sometime called "scale factor". That's because it is directly proportional to the cross-sectional area A of the diode. Thus doubling the junction area results in a diode with double the value of I_S and as the diode equation indicates, double the value of current i.
The value of I_S is of the order of 10^{-15} A however is a very strong function of temperature.

The parameter η can be determined from the exponential nature of the volt-ampere characteristic :

$$\log_{10} i_D = \log_{10} I_S + \frac{0.43\, v_D}{\eta V_T} \qquad \left\{ \log_{10} e \simeq 0.434 \right.$$

A plot of $\log i_D$ versus v_D results in a straight line of slope $\frac{0.434}{\eta V_T}$ from which η is obtained.

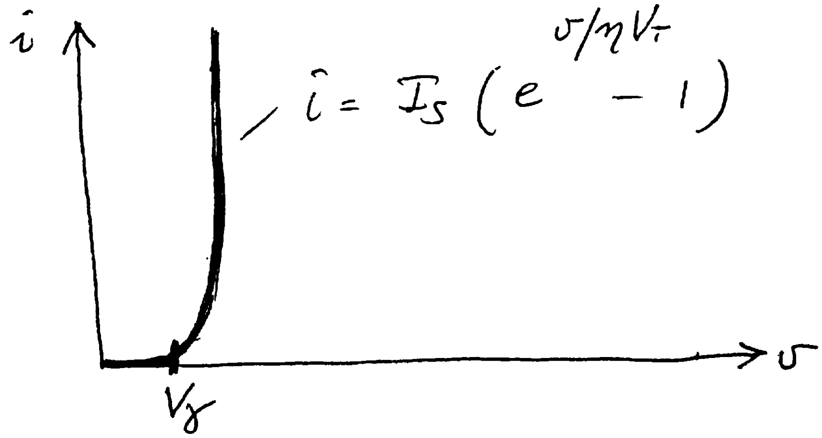
$1N4153 \rightarrow \eta \simeq 2$



$$\left\{ \begin{array}{ll} v > 0 & \text{Forward bias region} \\ -V_Z < v < 0 & \text{Reverse bias region} \\ v < -V_Z & \text{breakdown region} \end{array} \right.$$

diode operation regions

Forward region



A glance at the i-v characteristic in the forward region reveals that at room temperature the current is negligibly small for v smaller than about 0.6 V.
This value V_γ is usually referred as cut-in voltage. or threshold voltage.
Typical values of the threshold voltage (at room temperature) for different commonly used semiconductors are:

silicon $\rightarrow V_\gamma = 0.6\,V$
germanium $\rightarrow V_\gamma = 0.2\,V$
gallium arsenide $\rightarrow V_\gamma = 1.2\,V$

If we operate at room temperature and apply forward voltages we can reasonably assume that:

$$i = I_S\left(e^{v/\eta V_T} - 1\right) \underset{e^{v/\eta V_T} \gg 1}{\approx} I_S\, e^{v/\eta V_T}$$

$$10 = e^{v/V_T} \rightarrow v = (\ln 10)\, V_T = 2.3 \cdot 26\,mV \approx 52\,mV$$

for one order of magnitude

For voltages less than V_γ the curve can be approximated by a straight line of slope close to 0.
Since the slope ($\Delta i/\Delta v$) is the conductance that means that the conductance is very small in this region $\Rightarrow$ the resistance is very high

For voltages above V_γ the curve can be approximated by a straight line with a very large slope.
The conductance is therefore very large $\Rightarrow$ the equivalent resistance is very small

$$\Downarrow$$

$$g_d = \frac{di_D}{dv_D} = \frac{d}{dv_D}\left(I_S e^{v_D/\eta V_T} - I_S\right) = I_S \frac{d}{dv_D}\left(e^{v_D/\eta V_T}\right) =$$

$$= I_S \frac{e^{v_D/\eta V_T}}{\eta V_T}$$

$$\downarrow$$

$$g_d = \frac{di_D}{dv_D} = \frac{I_S e^{v_D/\eta V_T}}{\eta V_T} = \frac{I_S e^{v_D/\eta V_T} - I_S + I_S}{\eta V_T} =$$

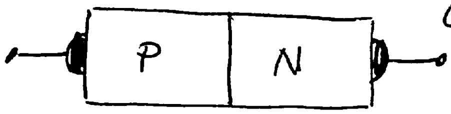
$$= \frac{i_D + I_S}{\eta V_T}$$

$$\underline{r_d = \frac{1}{g_d} = \frac{\eta V_T}{i_D + I_S}} \qquad \leftarrow \text{DYNAMIC RESISTANCE}$$

Usually not all the voltage applied at the diode terminals appears at the p-n junction there is a small drop due to the ohmic contact resistance between the metallic terminals and the semiconductor

$r_d + r_{contact}$ ← DIODE RESISTANCE

Anode



Cathode

Forward dynamic resistance

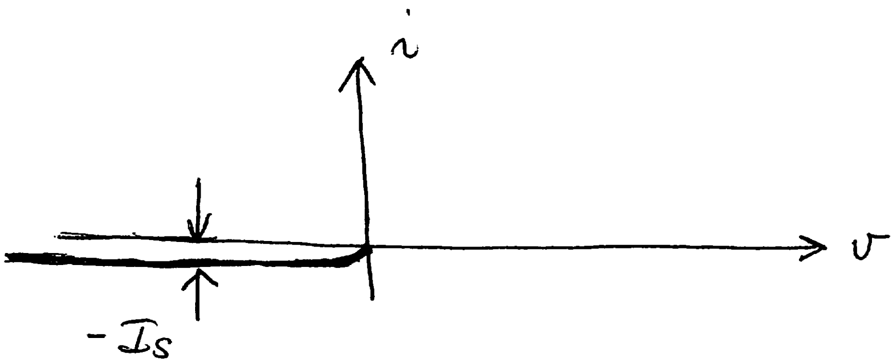
$$r_d = \frac{\eta V_T}{i_D + I_S} \underset{i_D \gg I_S}{\approx} \frac{\eta V_T}{i_D}$$

Reverse Region

The reverse bias region is entered when the diode voltage v is made negative.
If v is negative and few times larger than V_T in magnitude the exponential term becomes negligibly small compared to unity and we can reasonably write that:

$$i = I_S\left(e^{v/\eta V_T} - 1\right) \underset{\uparrow}{\approx} -I_S$$

$$e^{v/\eta V_T} \ll 1$$



This mean that:
The current in reverse direction is approximately constant and equal to I_S
→ this constancy is the reason behind the name <u>saturation</u> current.

$$r_d = \frac{\eta V_T}{i_D + I_S} \underset{\uparrow}{\approx} \infty$$

$i_D \approx -I_S$

← "Reverse "dynamic" resistance"

Breakdown region

The breakdown region is entered when the magnitude of the reverse voltage exceed a threshold value called the breakdown voltage or the zener knee voltage.

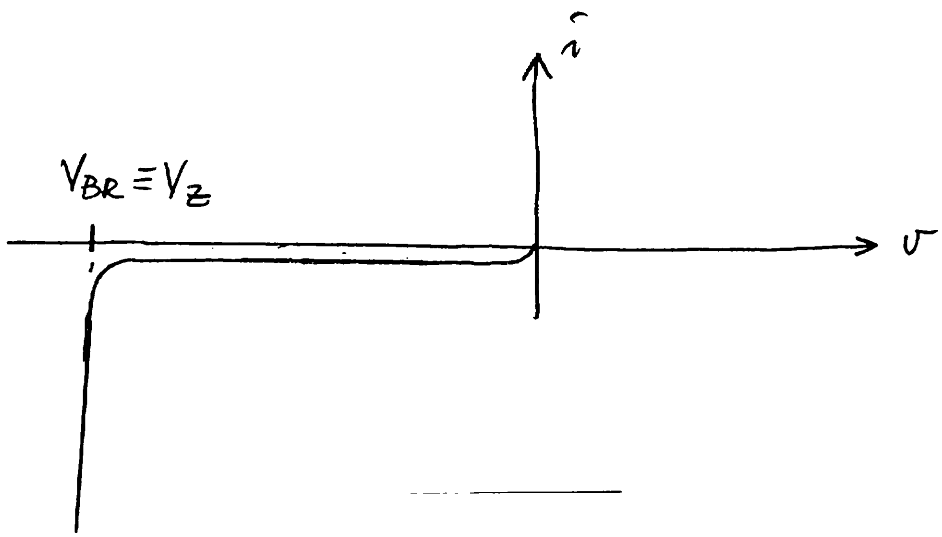
In the breakdown region the reverse current increases rapidly with a very small increase in the associated voltage.

The breakdown voltage is sometime called the peak inverse voltage (PIV) in manufacturer's specification sheets.

At breakdown there is an avalanche of electrons which flow across the junction with the result that the diode overheats.

The large current can destroy part of the diode (the portion that overheats).

Provided that the power dissipated in the diode is limited by external circuitry to a safe level (usually provided on the data sheets) breakdown won't be destructive.



TEMPERATURE EFFECTS

The diode characteristic has 2 terms, V_T and I_S which are heavily dependent on temperature.

$$V_T = \frac{kT}{q} = \frac{T}{11600}$$

$$\frac{I_S}{A} = q n_i^2 \left(\frac{D_p}{N_D L_p} + \frac{D_n}{N_A L_n} \right)$$ → which contains again V_T through Einstein relation

L = diffusion length = $\sqrt{D \cdot \tau}$ ← minory carrier lifetime (before recombination)

For silicon I_S increase approximately 15% per °C ($K_i = 0.15/°C$) → or in other terms I_S doubles every 5 °C

$$e^{0.15} \approx 1.15$$

$$\frac{1}{0.15} \simeq 7 \quad (*)$$

$$I_S(@T_2) = I_S(@T_1)\, 2^{(T_2-T_1)/5}$$

$$I_S(@T_2) = I_S(@T_1)\, e^{K_i (T_2-T_1)} = I_S(@T_1)\, e^{(T_2-T_1)/7}$$

(*) I have a value and I want it to become 1.15 times bigger every time I add a °C

↓

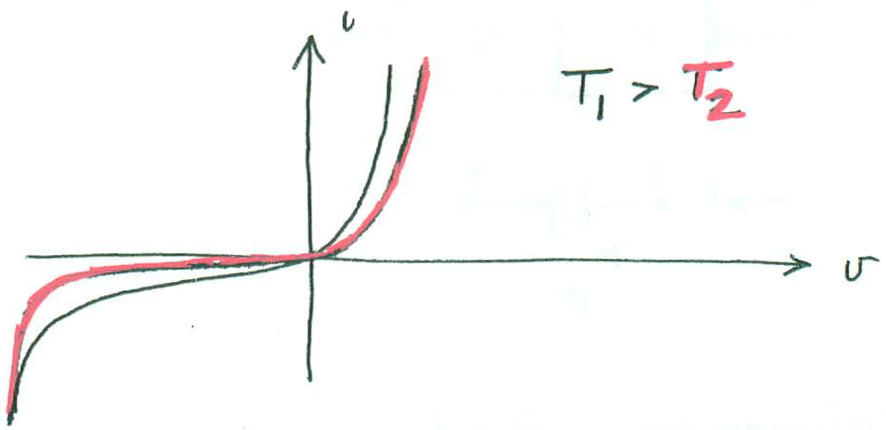
I need some mechanism that transform an addition in a multiplication

$$b^{x+y} = b^x \cdot b^y$$

$$b^{0.15 \cdot 1} = 1.15$$

$$b^{0.15 \cdot 2} = b^{0.15(1+1)} = b^{0.15} \cdot b^{0.15} = 1.15 \cdot 1.15$$

As T increases in order to mantain a constant value of I voltage must be reduced. For silicon $\frac{dV}{dT} \simeq -2.0$ mV/°C



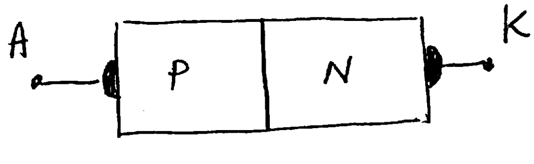
$$2^x = 1.15 \rightarrow \log_2 1.15 = x$$

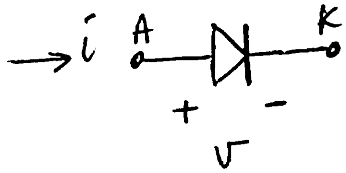
$$\log_2 \nu = \frac{\ln \nu}{\ln 2} = \frac{\log_{10} \nu}{\log_{10} 2}$$

$$\log_2 1.15 = \frac{\ln 1.15}{\ln 2} \simeq \frac{1}{5}$$

DIODE CONSTRUCTION

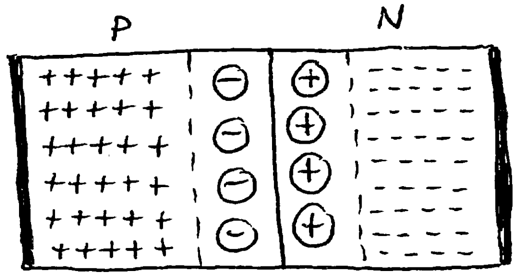
Semiconductor diodes are formed by a pn junction





The semiconductor materials commonly used are:
germanium, silicon, and gallium arsenide.
(GaAs is particularly used in microwave applications, photo-detection and laser diodes)

The unbiased p-n junction



+ majority holes

- majority free electrons

In the p-type material for simplicity are not shown the negative charges (in equal amount to the holes) associated with the ionized acceptor atoms and the minority free electrons generated by thermal agitation.

Similarly in the n-type material are not shown the positive charges associated with the ionized donor atoms and the minority holes generated by thermal agitation.

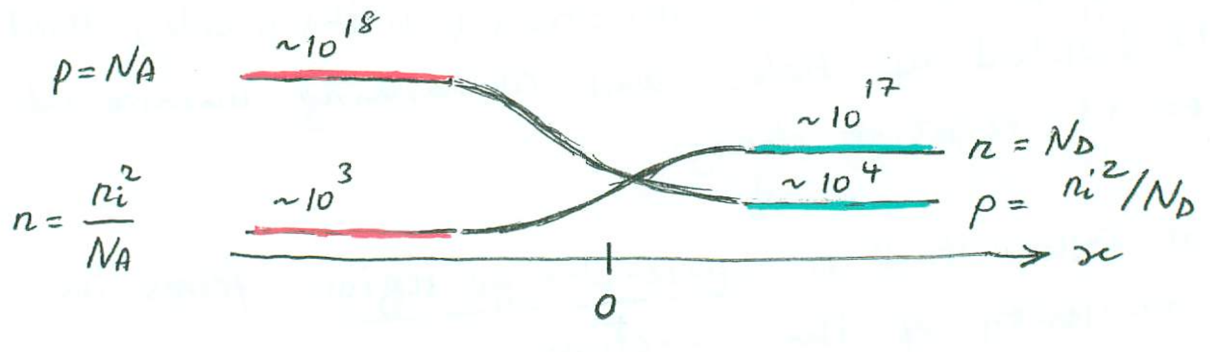
DIFFUSION CURRENT

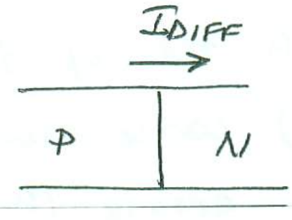
Because the concentration of holes is high in the p-region and low in the n-region, holes diffuse across the junction from the p side to the n side

Similarly the electrons diffuse across the junction from the n side to the p side

↓

These two components add together to form a diffusion current I_{DIFF} whose direction is from the p side to the n side





DEPLETION REGION

↓

The holes that diffuse across the junction into the n region will meet a high concentration of electrons so they quickly recombine disappearing from the scene.

The electrons that diffuse across the junction into the p region will quickly recombine with some of numerous holes present on the p side, thus disappear from the scene.

↓

As result there will be: a region close to the junction (on the n side) that is depleted of free electrons and contains uncovered bound positive charge

and

a region close to the junction (on the p side) that is depleted of holes and containing uncovered bound negative charge

↓

it follow that a space-charge region forms in proximity of the junction.

The charges on both side of the depletion region (= space charge region) cause an electric field to be established across the region.

(hence a potential difference result across the region with n side at positive voltage relative to the p side)

The resulting electric field opposes the diffusion of holes into the n region and electrons in the p region

↓

The voltage drop across the depletion region acts as a <u>barrier</u> that has to be overcome for holes to diffuse into the n region and electrons to diffuse into the p region.

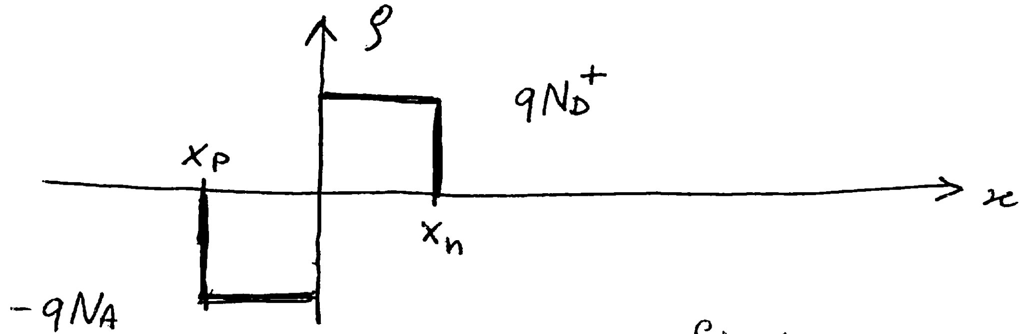


fig. 1

Gauss' Law :

$$\frac{dE_x}{dx} = \frac{\rho}{\varepsilon_{si}}$$

$$\rho = q\,(p - n + N_D - N_A)$$

The depletion region is free of electrons and holes

$N_D - N_A \gg p - n$

$\rho \cong q(N_D - N_A)$

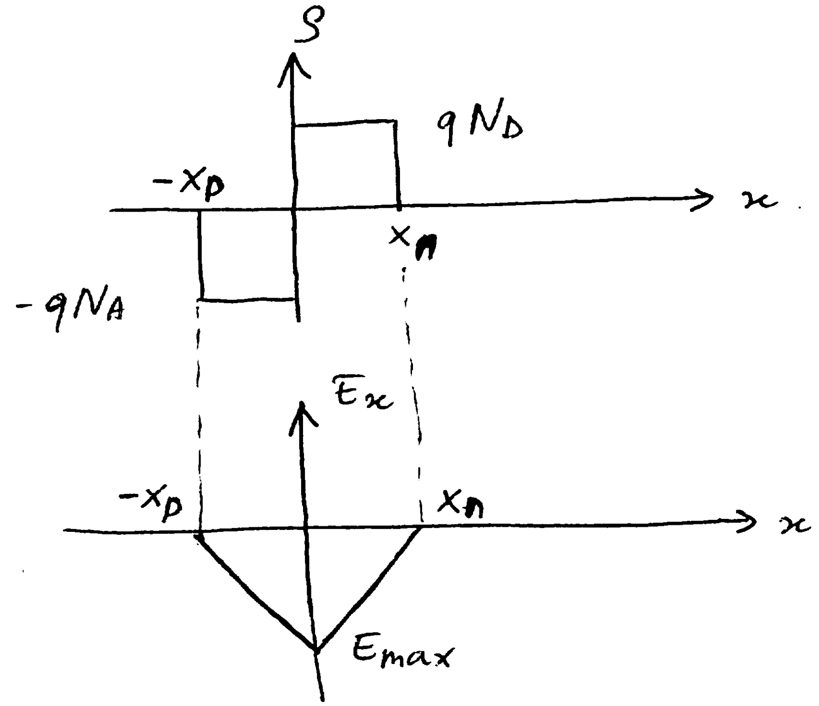
① on the p side $N_D = 0$

$$\frac{dE_x}{dx} = -\frac{qN_A}{\varepsilon_{si}}$$

② To the n side: $N_A = 0$

$$\frac{dE_x}{dx} = \frac{qN_D}{\varepsilon_{si}}$$

which plotted gives, fig.



① on the p-side

$$\int_{E_x(-x_p)}^{E_x(x)} dE_x = -\frac{qN_A}{\varepsilon_{si}} \int_{-x_p}^{x} dx$$

$$E_x(x) - \underbrace{E_x(-x_p)}_{0} = \frac{-qN_A}{\varepsilon_{si}}(x + x_p)$$

boundary of the space charge region

$$E_x(x) = -\frac{qN_A}{\varepsilon_{si}}(x + x_p)$$

$$E_x(0) = E_{max} = -\frac{qN_A}{\varepsilon_{si}} x_p$$

② on the n-side

$$\int_{E(x)}^{E(x_n)} dE_x = \frac{qN_D}{\varepsilon_{si}} \int_{x}^{x_n} dx$$

$$\underbrace{E(x_n)}_{= 0} - E_x(x) = \frac{qN_D}{\varepsilon_{si}} (x_n - x)$$

$$E_x(x) = \frac{-qN_D}{\varepsilon_{si}} (x_n - x)$$

$$E_x(0) = E_{max} = -\frac{qN_D}{\varepsilon_{si}} x_n$$

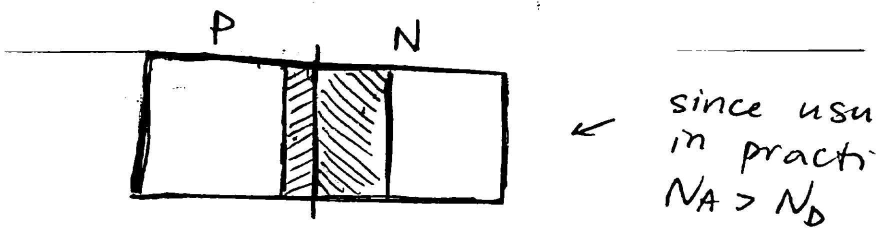
Thus:

$$E_x(0) = E_{max} = \frac{-qN_D}{\varepsilon_{si}} x_n = \frac{-qN_A}{\varepsilon_{si}} x_p$$

Field at the junction

$$\downarrow$$

$$N_D x_n = N_A x_p$$

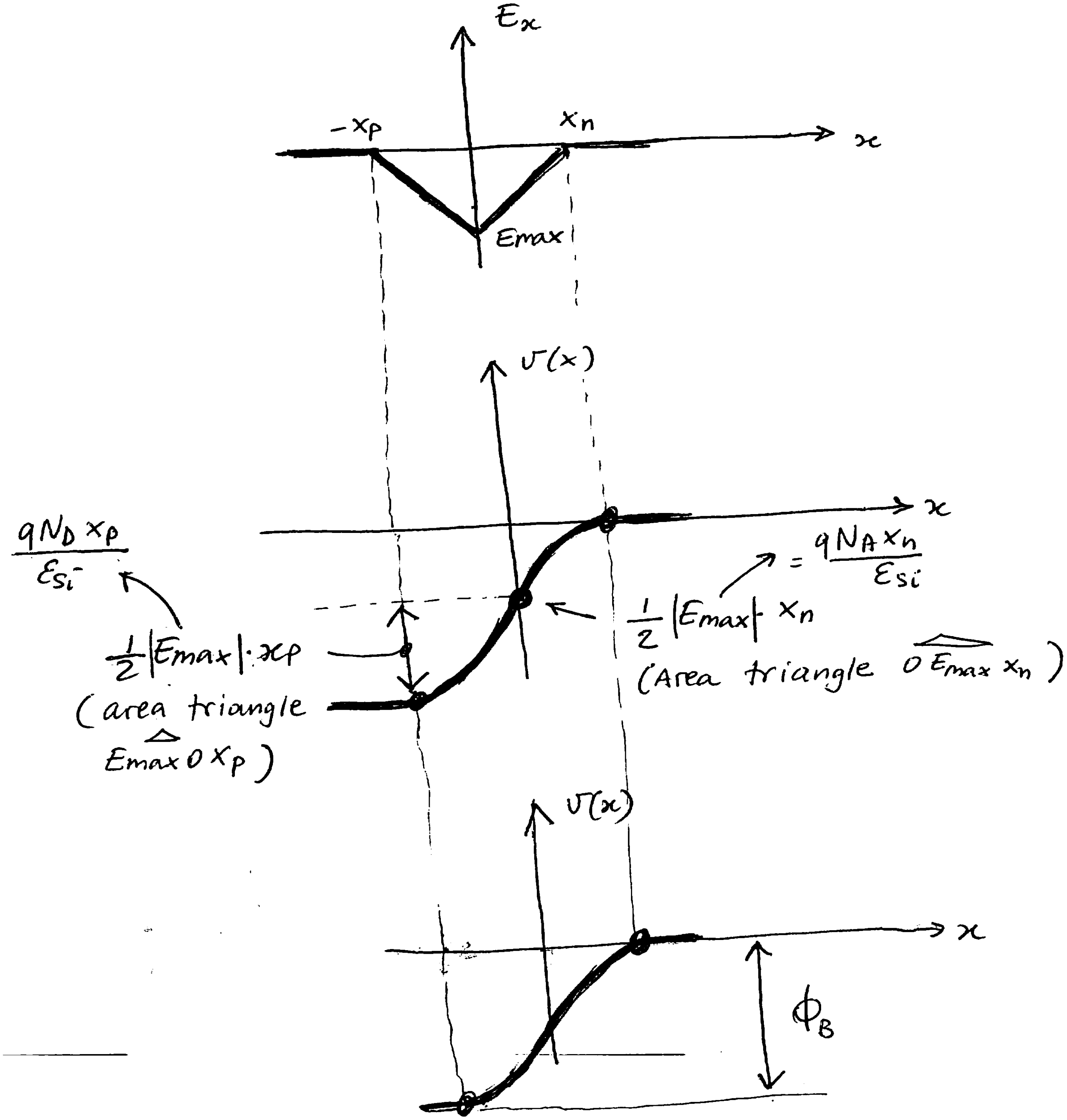


since usually in practice $N_A > N_D$

The width of the depletion region will not be the same on the 2 sides.

Recalling how potential and electric field are related

$$\frac{dv}{dx} = -E(x) \quad \rightarrow \quad dv = -E(x)\,dx$$



ϕ_B = junction built-in voltage =

$$= \frac{1}{2} \frac{q N_D}{\varepsilon_{si}} x_n^2 + \frac{1}{2} \frac{q N_A}{\varepsilon_{si}} x_p^2$$

Recalling that:

$$N_D x_n = N_A x_p \rightarrow \begin{cases} x_p = (N_D/N_A)\, x_n \\ x_n = \dfrac{N_A}{N_D}\, x_p \end{cases}$$

$$x_{dep} = x_p + x_n = \left(\frac{N_A}{N_D} + 1 \right) x_p = \left(\frac{N_D}{N_A} + 1 \right) x_n$$

$$x_p = x_{dep} \frac{N_D}{N_D + N_A}$$

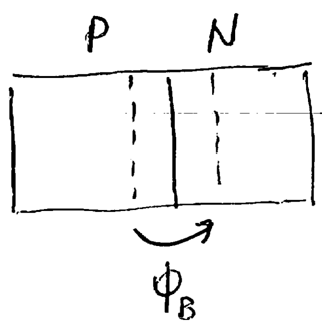
$$x_n = x_{dep} \frac{N_A}{N_D + N_A}$$

$$\Downarrow$$

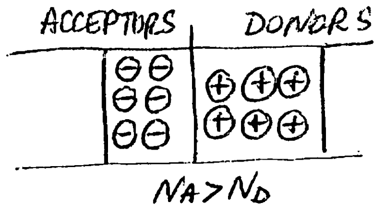
$$x_{dep} = \sqrt{\frac{2 \varepsilon_{si}}{q} \cdot \frac{N_A + N_D}{N_A N_D} \phi_B}$$

WIDTH OF THE DEPLETION REGION

$$|\mathcal{E}_{max}| = \sqrt{\frac{q}{2 \varepsilon_{si}} \frac{N_A N_D}{N_A + N_D} \cdot \phi_B}$$

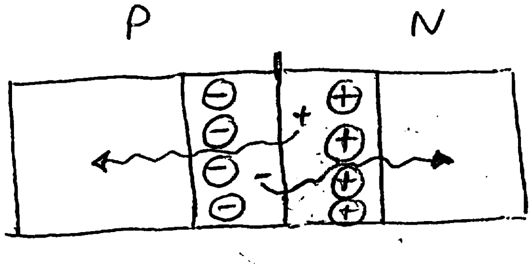


As already noticed since usually the doping levels are not equal in the p and n materials the width of the depletion region will not be the same on the two sides
→ in order to uncover the same amount of charge the depletion layer will extend deeper into the more ligthly doped material.



Equilibrium
↓
Across the junction:
In addition to the current component I_{DIFF} due to the majority carrier diffusion there is as well a component due to minority carrier (I_S).

Specifically some of the thermally generated holes in the n material travel through the n material to the edge of the depletion region. There they experience the electric field in the depletion region, which , them across that region into the p side.

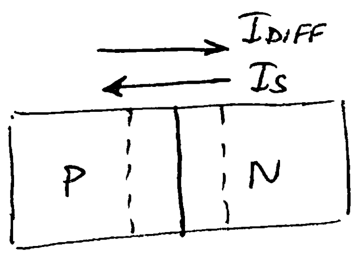


Similarly some of the thermally generated free electrons in the p material reach the edge of the depletion region and get swept by the electric field in the depletion region across that region into the n-side.

These two current components electrons moved by drift from p to N
and holes moved by drift from N to P
add together to form a drift current I_S, whose direction is from the N side to the P side

Under the open-circuit condition no current exists thus the two opposite currents across the junction must be equal

$$I_{DIFF} = I_S$$



$$E_x = -\frac{dv}{dx}$$

$$J = J_p = J_n = 0$$

$$\begin{cases} J_p = q\mu_p p E_x - q D_p \dfrac{dp}{dx} = 0 & (1) \\ J_n = q\mu_n n E_x + q D_p \dfrac{dn}{dx} = 0 & (2) \end{cases}$$

solution of the differential eq. (1) is:

solution of the differential eq. (2) is:

$$p(x) = p(x')\, e^{-\frac{v(x)-v(x')}{KT/q}}$$

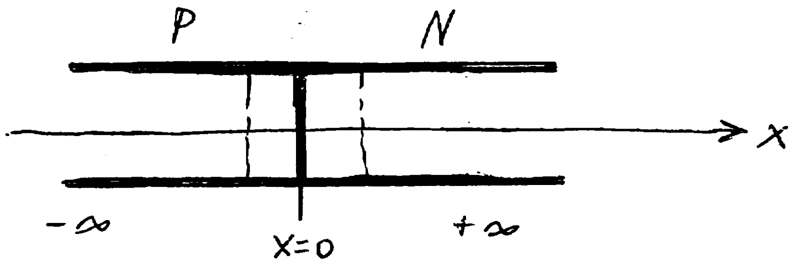
$$n(x) = n(x')\, e^{\frac{v(x)-v(x')}{KT/q}}$$

SEE HANDOUT #7

$$\Downarrow$$

$$p(\infty) = p(-\infty)\, e^{-\frac{\phi_B}{KT/q}}$$

$$n(\infty) = n(-\infty)\, e^{+\frac{\phi_B}{KT/q}} \qquad (*)$$



$$p(-\infty) \approx N_A \qquad\qquad n(+\infty) \cong N_D$$

$$n(-\infty) = \frac{n_i^2}{N_A} \qquad\qquad p(+\infty) = \frac{n_i^2}{N_D}$$

$\Downarrow$ substituting in $(*)$

$$N_D = \frac{n_i^2}{N_A}\, e^{\frac{\phi_B}{KT/q}} \rightarrow \frac{N_D N_A}{n_i^2} = e^{\frac{\phi_B}{KT/q}} \rightarrow$$

$$\rightarrow \frac{\phi_B}{KT/q} = \ln \frac{N_D N_A}{n_i^2}$$

$\downarrow$

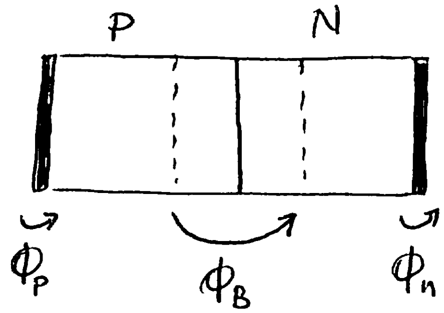
$$\boxed{\phi_B = \frac{KT}{q} \ln \frac{N_D N_A}{n_i^2}}$$

BUILT IN VOLTAGE

Typically for silicon at room temperature ϕ_B is in the range $0.6 \div 0.8$ V $\longrightarrow$ $26\,\text{mV} \cdot \ln \frac{10^{18} \cdot 10^{17}}{10^{21}} \simeq 0.83$

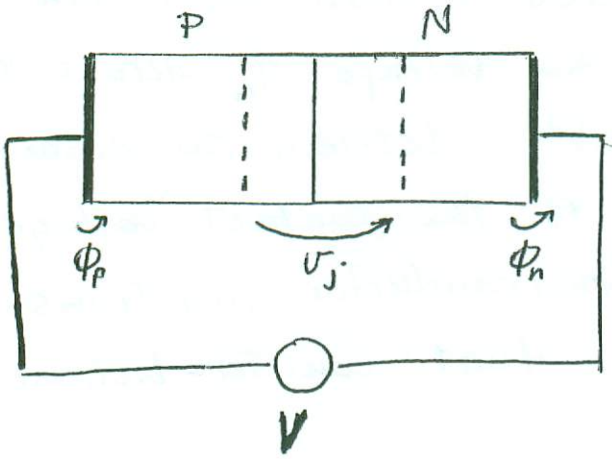
NOTE ϕ_B is not V_γ !!!

When the p-n junction terminals are left open-circuited the voltage measured between them will be zero! That means that the voltage ϕ_B across the depletion region does not appear between the diode terminals. This is because of the contact voltages existing at the metal-semiconductor junctions at the diode terminals that counter-balance the barrier voltage.

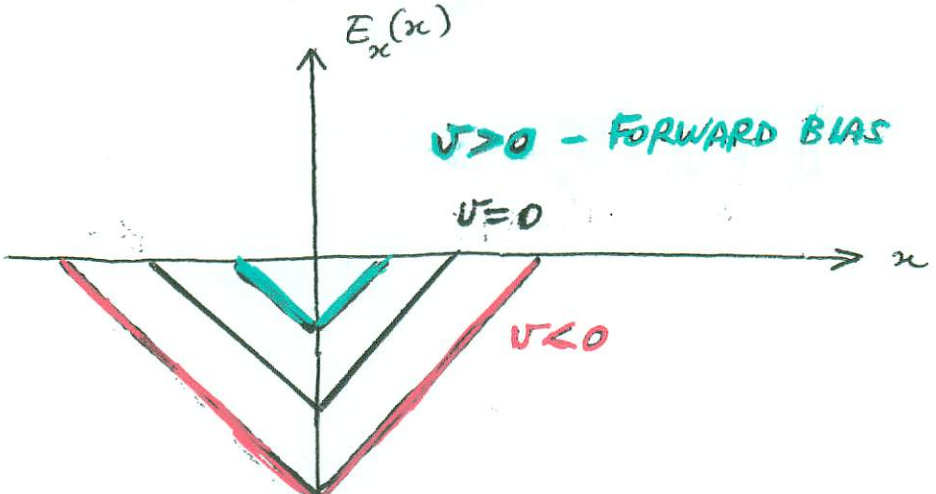


$$\phi_P + \phi_B + \phi_n = 0$$

- The BIASED P-N Junction

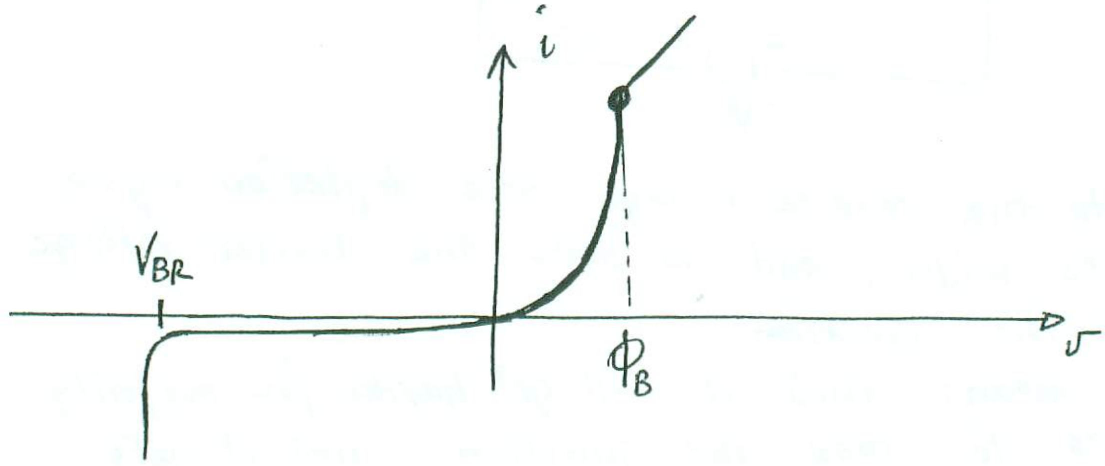


$$v + \underbrace{\phi_p + \phi_n}_{-\phi_B} + v_j = 0 \quad \rightarrow \quad v_j = \phi_B - v$$



If I keep decreasing the voltage too much eventually I'll reach a breakdown situation (depletion region can't get bigger than the length of the semiconductor bar !!!)

If I keep increasing the voltage eventually the depletion region will disappear (when $v = \phi_B$)
As v becomes comparable with ϕ_B the pn junction behaves like a sort of resistor (the current is governed by the ohmic-contact and the semiconductor-bulk resistance)

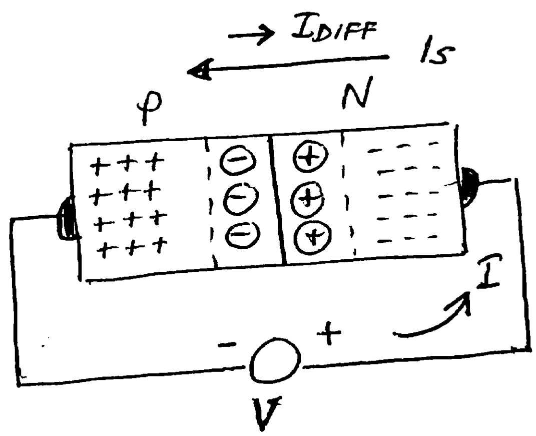


$$\phi_B = \frac{KT}{q} \ln \frac{N_A N_D}{n_i^2}$$

$$E_{max} = \sqrt{\frac{2q}{\varepsilon_{Si}} \frac{N_A N_D}{N_A + N_D} (\phi_B - v)}$$

$$x_{dep} = \sqrt{\frac{2\varepsilon_{Si}}{q} \frac{N_A + N_D}{N_A N_D} (\phi_B - v)}$$

REVERSE BIASED PN-Junction $v < 0$



Due to the reverse voltage the depletion region becomes wider, and so does the barrier voltage across the junction
This means that it will get harder for majority carriers to cross the junction, but it gets easier for the minority carriers to be swept (drifted) from one side to the other.

$$I = I_S - I_D \approx I_S$$

I_S dominates I_{DIFF}

If we keep decreasing the voltage the depletion region becomes wider and wider until at a certain point there is a new phenomenon that will set ~~in and~~ supply the charge ~~carriers~~ !
This is called breakdown !!

There are two mechanism that contributes to the breakdown:

* zener effect
* avalanche effect

→ If $|V_{BR}| > 5V$ usually the breakdown is due to zener effect.

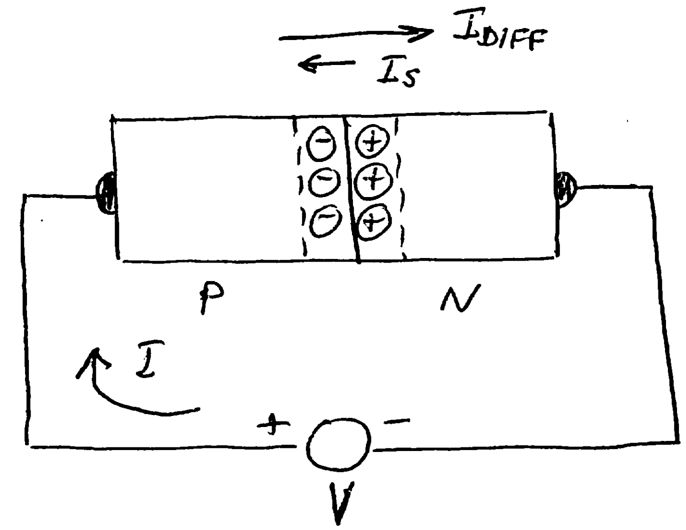
→ If $|V_{BR}| > 7V$ usually the breakdown is due to avalanche effect.

→ For junctions that breaks down between 5 and 7 V usually we have a combination of the two.

ZENER BREAKDOWN → occurs when the electric field in the depletion layer is strong enough to break covalent bonds and generate electron-hole pairs

AVALANCHE BREAKDOWN → the minority carriers that cross the junction under the influence of the electric field gain enough kinetic energy to be able to break covalent bonds in atoms with which they collide.

FORWARD BIASED PN-junction $v > 0$



Due to the forward voltage the depletion region shrink, and so does the voltage barrier across the junction

This means that it will be easier for the majority carriers to overcome the barrier and therefore cross the junction
but it is tougher for the minority carriers to be drifted all the way from one side to the other.

$$I = I_{DIFF}\ I_S \approx I_{DIFF}$$

I_{DIFF} dominates I_S

Current - voltage relationship

$$J = \frac{I}{A} \rightarrow I = A \cdot J$$

There are two contributes: one due to the minority carriers and the other due to the majority carriers.
The contribute due to the majority carriers is the diffusion current
The contribute due to the minority carrier is the drift current
It can be shown that:

$$J = \left(\frac{q D_p n_i^2}{L_p N_D} + \frac{q D_n n_i^2}{L_n N_A} \right)\left(e^{V/V_T} - 1\right)$$

$\frac{n_i^2}{N_D}$ = concentration of holes in the **N** region (→ minory carriers)

$\frac{n_i^2}{N_A}$ = concentration of the electrons in the P region (→ minority carriers)

D_p diffusion constant for holes in the N-type silicon

L_p diffusion length of holes in the N-type silicon

$$L_p = \sqrt{D_p \cdot \tau_p}$$

τ_p average time it takes for a hole to into the n-type silicon to recombine with a majority electron

...

$$I = A \cdot J = \underbrace{A q n_i^2 \left(\frac{D_p}{L_p N_D} + \frac{D_n}{L_n N_A} \right)}_{\substack{\parallel \\ I_S}} \left(e^{V/V_T} - 1 \right)$$

Diode Capacitances

Depletion capacitance

The depletion layer stores a charge of equal amount on each side of the junction. → it forms a capacitance !!

$$q_j = q_N = q N_D x_n A = q_P = q N_A x_p A$$

$$\downarrow$$

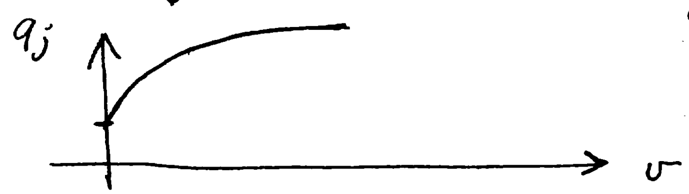
$$q_j = q \frac{N_A N_D}{N_A + N_D} A \, x_{dep}$$

and recalling what we found some time ago:

$$x_{dep} = \sqrt{\frac{2 \varepsilon_{si}}{q} \frac{N_A + N_D}{N_A \cdot N_D} (\phi_B - v)}$$

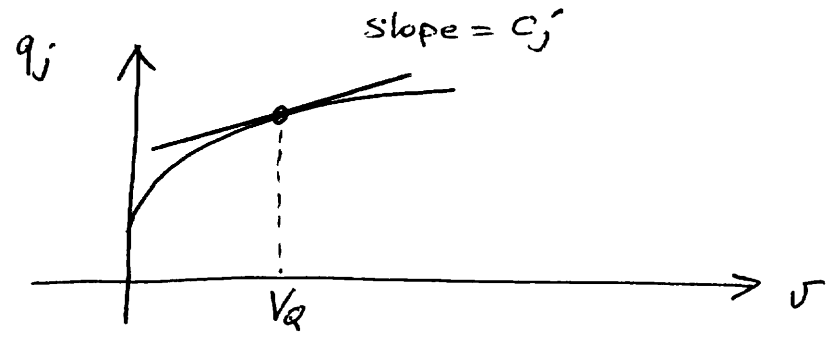
$q_j - v$ relationship is not linear !! so this is a non linear capacitor !!

$$\downarrow$$



$q_j \propto \sqrt{v}$

Obviously a linear-capacitance approximation can be done if the device is biased and signal swing around the bias point is small!



$$C_j = \left. \frac{dq_j}{dv} \right|_{v = V_Q}$$

or using the familiar parallel-plate capacitor formula:

$$C_j = \frac{\varepsilon_S \cdot A}{X_{dep}} = \frac{\varepsilon_S \cdot A}{\sqrt{\frac{2\varepsilon_S}{q} \cdot \frac{N_A + N_D}{N_A \cdot N_D} (\phi_B - v)}} =$$

$$= \frac{A}{\sqrt{\frac{2}{q\varepsilon_S} \left(\frac{N_A + N_D}{N_A \cdot N_D} \right) (\phi_B - v)}} =$$

$$= \frac{A}{\sqrt{\frac{2}{q\varepsilon_S} \left(\frac{N_A + N_D}{N_A \cdot N_D} \right) (\phi_B - v) \frac{\phi_B}{\phi_B}}} =$$

$$= \frac{A}{\sqrt{\frac{2}{q\varepsilon_S} \left(\frac{N_A + N_D}{N_A \cdot N_D} \right) \left(1 - \frac{v}{\phi_B}\right) \phi_B}} =$$

$$C_{j0} \triangleq \frac{A}{\sqrt{\frac{2}{q\varepsilon_s}\left(\frac{N_A + N_D}{N_A \cdot N_D}\right)\phi_B}}$$

Therefore we can write

$$C_j = \frac{C_{j0}}{\sqrt{1 - \frac{v}{\phi_B}}}$$

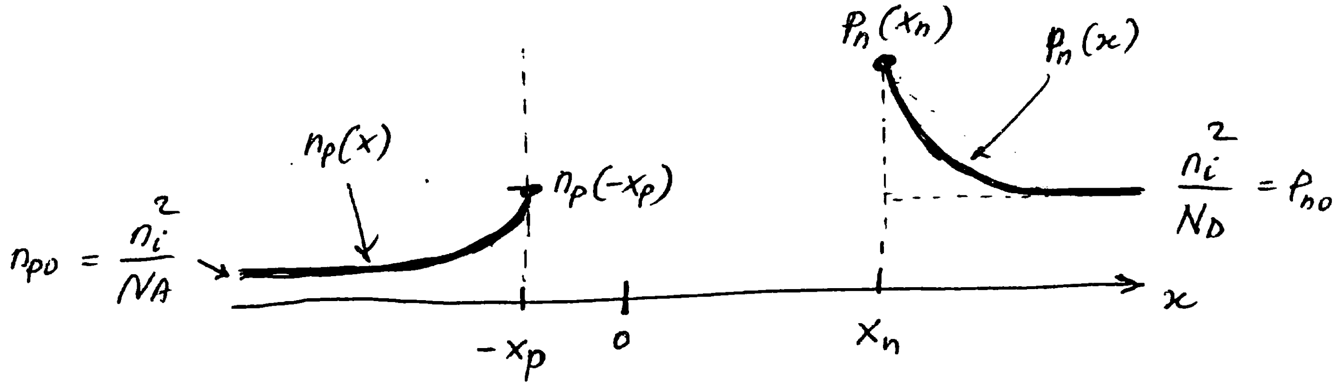
The analysis is based on the assumption that the carrier concentration change abruptly at the junction boundary.
More in general

$$C_j = \frac{C_{j0}}{\left(1 - \frac{v}{\phi_B}\right)^{m_j}}$$

where m_j is called GRADING COEFFICIENT and its value ranges from $\frac{1}{3}$ to $\frac{1}{2}$ depending on the manner in which the concentration changes from the p to N side of the junction.

Diffusion Capacitance

In the vicinity of the junction on the N side we have greater hole concentration than normally exists (→ because of diffusion).
This excess hole density can be considered as charge storage in the neighborhood of the junction.
Similar statement apply to electrons which diffuse into the p region



The excess hole charge stored in the n region is given by:

$$Q_p = A \cdot q \cdot [\, p_n(x_n) - p_{no} \,] \cdot L_p =$$

$$= A \cdot J_p \frac{L_p^2}{D_p} = I_p \cdot \frac{L_p^2}{D_p} =$$

$$= I_p \cdot \tau_p$$

$$p_n(x_n) = p_{no}\, e^{V/V_T}$$

$$J_p = q \frac{D_p}{L_p} p_{no} \left(e^{V/V_T} - 1\right)$$

$$p_{no}\left(e^{V/V_T} - 1\right) = \frac{J_p L_p}{D_p \cdot q}$$

$$L_p = \sqrt{D_p \tau_p}$$

$$L_p^2 = D_p \tau_p$$

In a similar way:

$$Q_n = I_n \cdot \tau_n$$

The total excess minority-carrier charge is:

$$Q = Q_p + Q_n = I_p \tau_p + I_n \tau_n$$

since the diode current $I = I_p + I_n$ we can express the excess charge as:

$$Q = \tau_T \cdot I$$

where τ_T is called diode mean <u>transit time</u>.
Pratically since usually $N_A >> N_D \rightarrow I_p >> I_n$
then $I \approx I_p$ and $Q_p >> Q_n \rightarrow Q \approx Q_p$
and thus $\tau_T \approx \tau_p$
This solve the problem of finding out τ_T!

For small changes around a bias point:

$$C_D = \left.\frac{dQ}{dV}\right|_{V=V_Q}$$

$$\downarrow$$

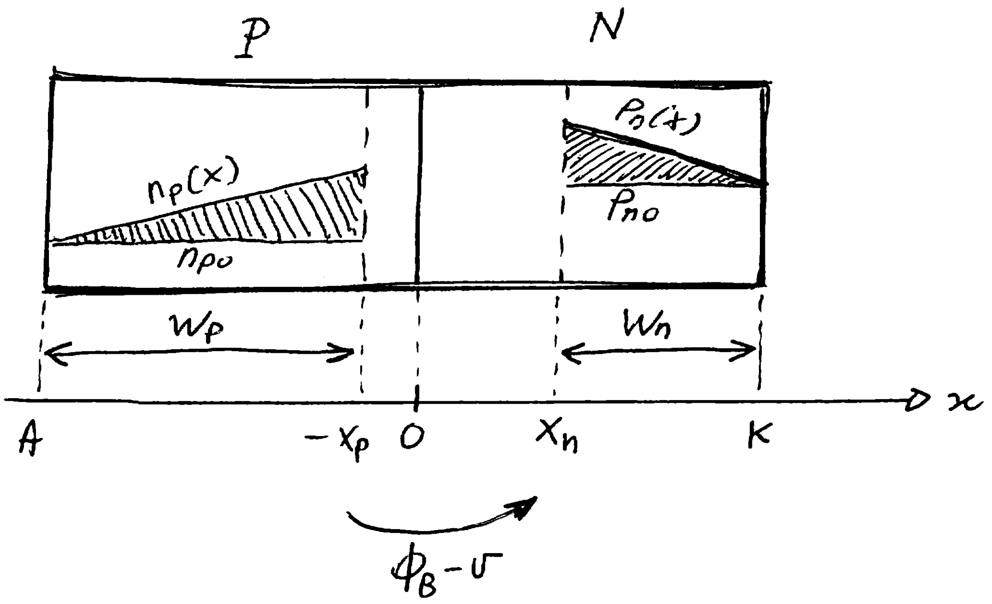
$$C_D = \frac{d(\tau_T I)}{dV} = \tau_T \left.\left(\frac{dI}{dV}\right)\right|_{V=V_Q} =$$

$$= \tau_T \cdot g_{diode}$$

and we found some time ago that:

$$g_{diode} = \frac{I_{DIODE} + I_S}{V_T}$$

HOW DOES THE CURRENT-VOLTAGE REL. COME OUT ?



There is electric field only in the space-charge region (between $-x_p$ and x_n)

↓

Therefore the region from A to $-x_p$ and from x_n to K are quasi neutral !!

The boundary of the quasi-neutral p-region is at $-x_p$ and the hole density there must be equal at equilibrium as well as under bias

The same is true for the electron density at the boundary of the quasi-neutral n-region at x_n

↓ can be shown (see Muller-Kamins p. 234)

$$n_p(-x_p) = n_{po}\, e^{v/V_T}$$

$$p_n(x_n) = p_{no}\, e^{v/V_T}$$

The current that flows through any transverse section x must be constant and has the contribute of both holes and electrons:

$$I_{DIODE} = I_n + I_p$$

$$I_n = I_{n,drift} + I_{n,diff}$$

$$I_p = I_{p,drift} + I_{p,diff}$$

If we consider the quasi-neutral regions than since there is no field in those regions there will be no drift:

$$I_n \approx I_{n,diff} \quad \text{(in the neutral region } W_n\text{)}$$

$$I_p \approx I_{p,diff} \quad \text{(in the neutral region } W_p\text{)}$$

On this basis, the most suitable transverse sections for the evaluation of the total current I_{DIODE} are those at the boundary of the depletion layer $\rightarrow$ $x = -x_p$ or $x = x_n$

$$I_{DIODE} = I_n(-x_p) + I_p(-x_p)$$

since the flux of the carriers is constant in the depletion layer ($\rightarrow$ simplifying assumption !!)

$$I_p(-x_p) = I_p(x_n)$$

$$\downarrow$$

$$I_{DIODE} = I_n(-x_p) + I_p(x_n)$$

currents of the minority carriers

$$\downarrow$$

$$I_{DIODE} = I_{n,diff}(-x_p) + I_{p,diff}(x_n)$$

The diffusion currents are $\propto$ to the gradient of the distribution of the carriers:

① $I_{n,diff}(x) \quad qAD_n \dfrac{dn_p(x)}{dx}$ (for electrons)

② $I_{p,diff}(x) = -qAD_p \dfrac{dp_n(x)}{dx}$ (for holes)

$\downarrow$ I assume linear distributions !! $\searrow$ SHORT DIODE

$$\frac{dn_p}{dx} = \frac{n_p(-x_p) - n_p(A)}{W_p} = \frac{n_{po}e^{v/V_T} - n_{po}}{W_p}$$

$$\frac{dp_n}{dx} = \frac{-p_n(x_n) + p_n(K)}{W_n} = \frac{-p_{no}e^{v/V_T} + p_{no}}{W_p}$$

$$\downarrow$$

$$\frac{dn_p}{dx} = \frac{n_i^2}{N_A \cdot W_p}\left(e^{v/V_T} - 1\right)$$

$\uparrow$ $n_{po} = n_i^2/N_A$

$$\frac{dp_n}{dx} = \frac{-n_i^2}{N_D \cdot W_n}\left(e^{v/V_T} - 1\right)$$

$\uparrow$ $p_{no} = n_i^2/N_D$

Substituting in ① and ②

① $$I_{n,diff}(-x_p) = \frac{n_i^2}{N_A W_p} qAD_n \left(e^{v/V_T} - 1\right)$$

② $$I_{p,diff}(x_n) = \frac{n_i^2}{N_D W_n} qAD_p \left(e^{v/V_T} - 1\right)$$

$$\Downarrow$$

$$I_{DIODE} = \underbrace{qAn_i^2\left(\frac{D_n}{N_A W_p} + \frac{D_p}{N_D W_n}\right)}_{\substack{\parallel \\ I_S}} \left(e^{v/V_T} - 1\right)$$

In the case of a LONG DIODE:

$$I_{DIODE} = \underbrace{qAn_i^2\left(\frac{D_n}{N_A L_n} + \frac{D_p}{N_D L_p}\right)}_{\substack{\parallel \\ I_S}} \left(e^{v/V_T} - 1\right)$$

A the minority carriers will recombine before reaching the diode terminals !!!

The diode as a circuit element

If one is doing a pencil and paper **design** of a relatively complex circuit, rapid circuit analysis is a necessity

↓

we want to quickly evaluate various possibilities before committing to a suitable circuit topology

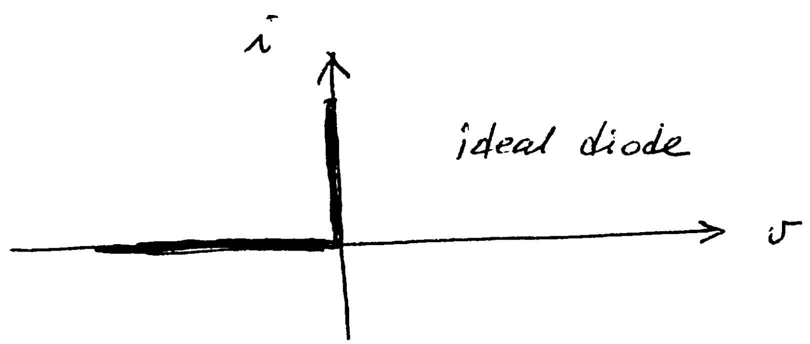
More accurate analysis can be postponed until the almost final design is obtained → possibly with the aid of a computer circuit-analysis program (e.g. SPICE), then

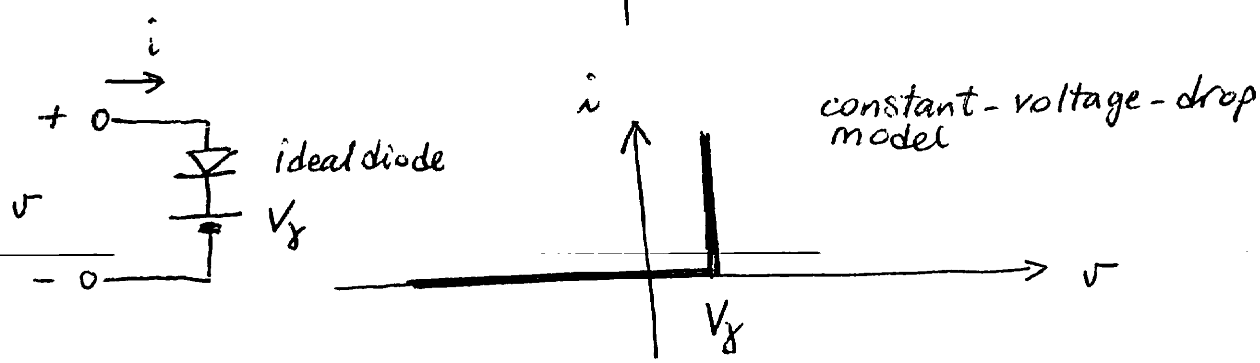
The accurate can be used to further refine (fine-tune) the design.

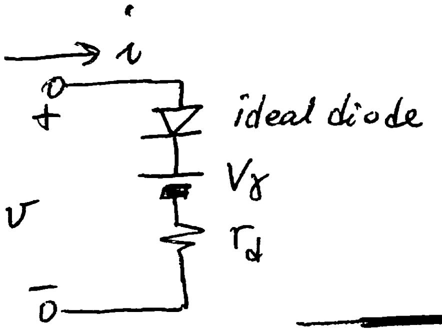
↓

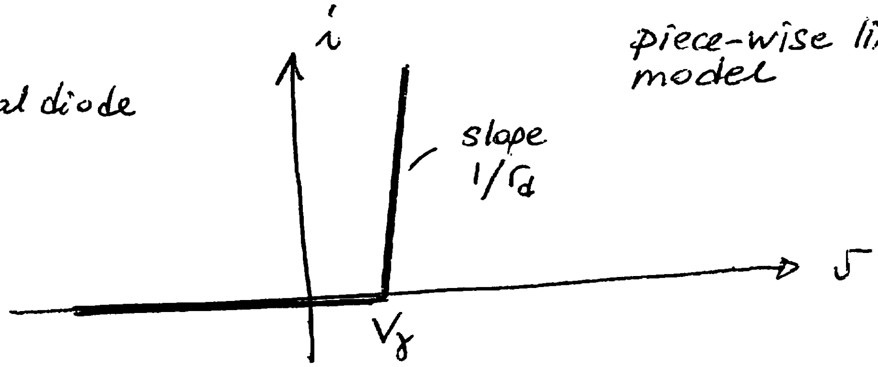
The issue is of finding an appropriate compromise between accuracy and complexity.

Different simplified models of the diode are commonly used.



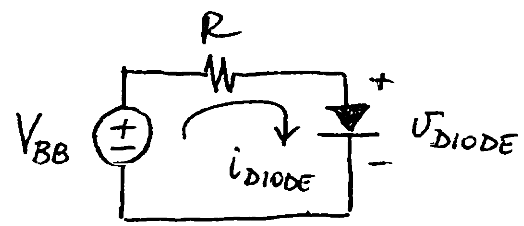


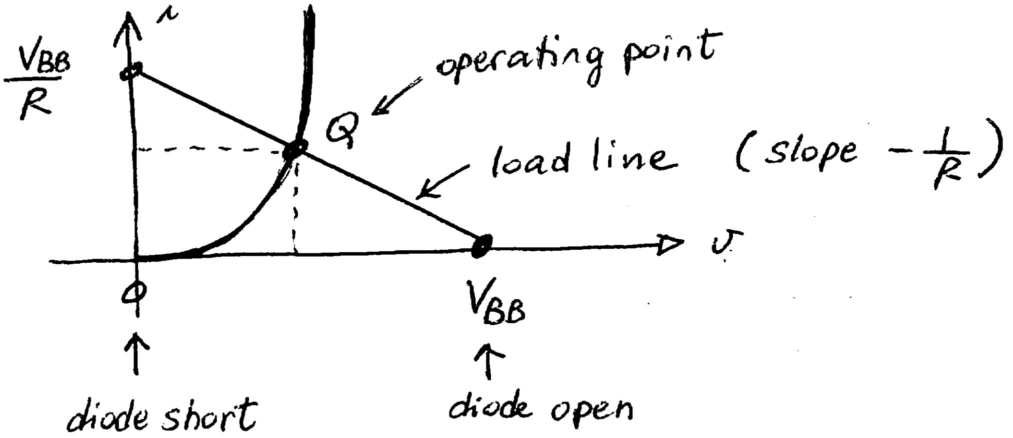




piece-wise linear model

The Load Line concept



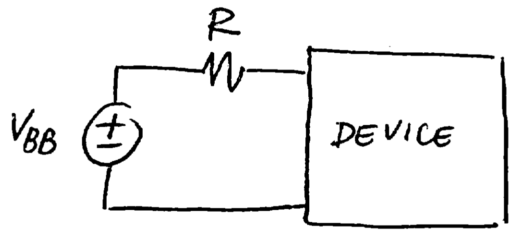


$$V_{BB} = R\, i_{DIODE} + v_{DIODE} \longrightarrow i_{DIODE} = \underbrace{\frac{V_{BB}}{R} - \frac{v_{DIODE}}{R}}$$

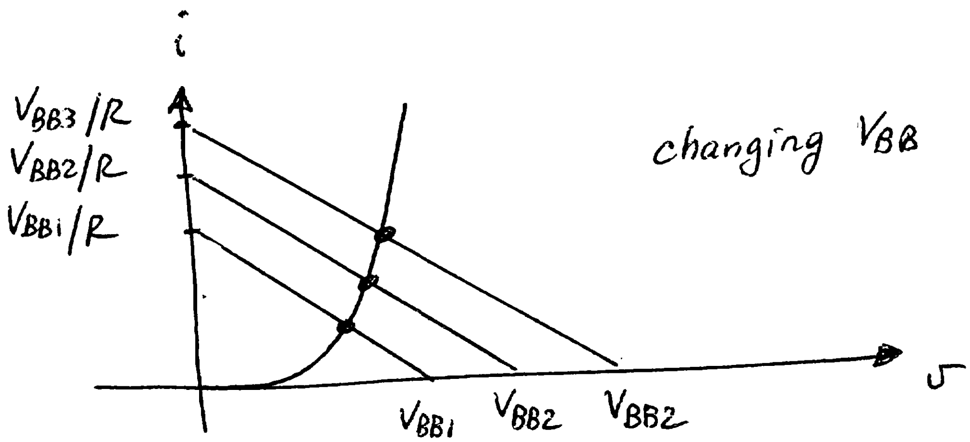
this is the equation of a line:

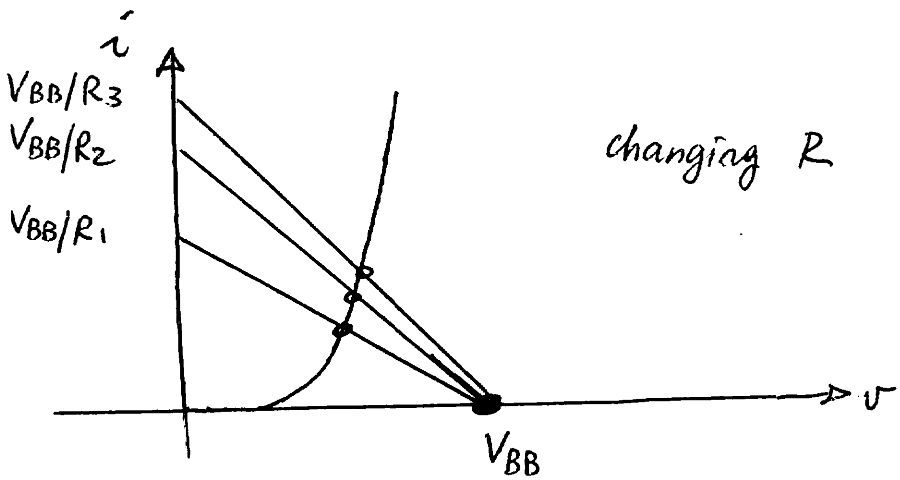
$y = a - bx$

The intersection of the load line with volt-ampere carachteristic of the diode gives the operating values of voltage and current in the circuit.



This is a general concept does not apply just for diodes.



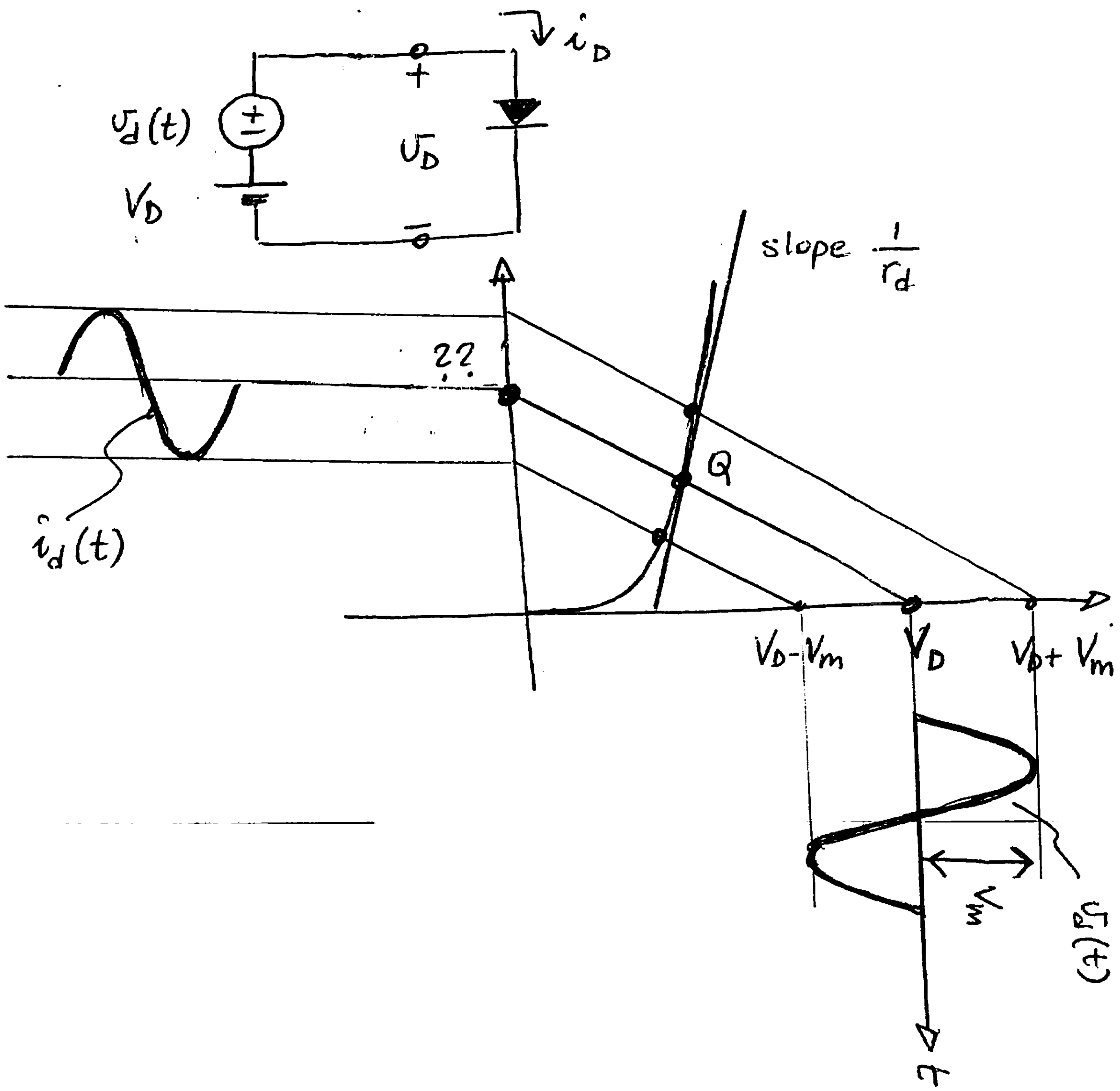


Small signal model

So far we have been mainly focused on using the ON-OFF behavior of the diode.
In these applications usually the applied signal is relatively large so the models seen so far are more than adequate!

↓

There are other application where the diode is biased to operate at a certain point (V_{DQ}, I_{DQ}) on the i-v characteristic and a small ac signal is superimposed on the dc quantities



When the signal $v_d(t)$ is absent the diode voltage is equal to:

$$v_D = V_D \leftarrow V_{DQ}$$

and the diode will conduct a dc current

$$i_D = I_D \cong I_S e^{V_D/\eta V_T} \leftarrow I_{DQ}$$

when the signal $v_d(t)$ is applied the total istantaneous diode voltage will be given by

$$v_D(t) = V_D + v_d(t)$$

correspondingly the Total istantaneous current $i_D(t)$ will be:

$$i_D(t) \cong I_S e^{v_D/\eta V_T} =$$

$$= I_S e^{(V_D/\eta V_T)} e^{v_d/\eta V_T} =$$

$$= I_{DQ} \cdot e^{v_d/\eta V_T}$$

$$f(x) = f(x_0) + f'(x_0)(x - x_0) + \frac{f''(x_0)(x - x_0)^2}{2!} + \dots$$

TAYLOR
EXPANSION

$$e^{ax} = e^{ax_0} + a e^{ax_0}(x - x_0) + \dots =$$

$x_0 = 0$

$$= 1 + ax + \dots$$

Therefore if $\frac{v_d}{\eta V_T} \approx 0$ (as rule of thumb take amplitudes smaller than 10 mV)

$$i_D(t) \cong I_{DQ} \cdot \left(1 + \frac{v_d}{\eta V_T}\right) \rightarrow$$

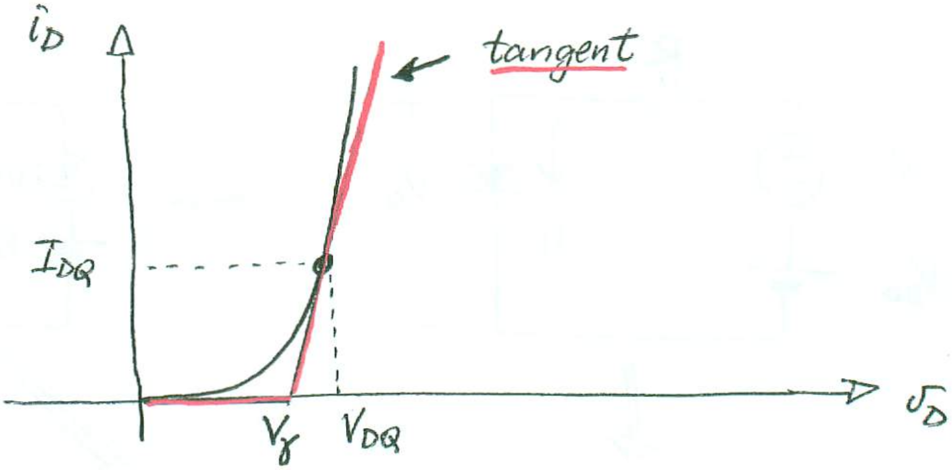
$$\rightarrow i_D(t) = \underbrace{I_{DQ}}_{I_D} + \underbrace{I_{DQ}\frac{v_d(t)}{\eta V_T}}_{i_d(t)}$$

$$\boxed{i_d(t) = \frac{I_{DQ}}{\eta V_T} v_d(t)}$$

$$\frac{\eta V_T}{I_{DQ}} \triangleq r_d \leftarrow$$ diode small signal resistance or incremental resistance

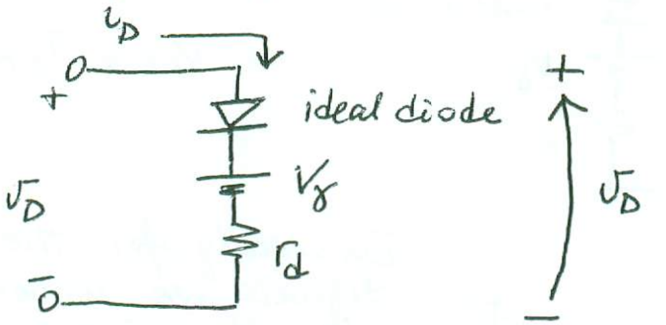
FORWARD BIASED DIODE !!!

If we look at what we have done from a graphical perspective;



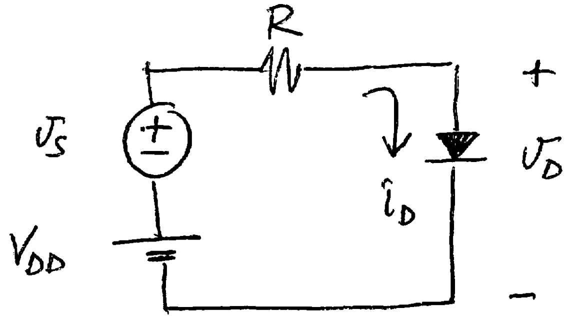
$$i_D = \frac{1}{r_d}\left(v_D - V_\gamma\right)$$ ← equation of the tangent to the i-v diode curve at (V_{DQ}, I_{DQ})

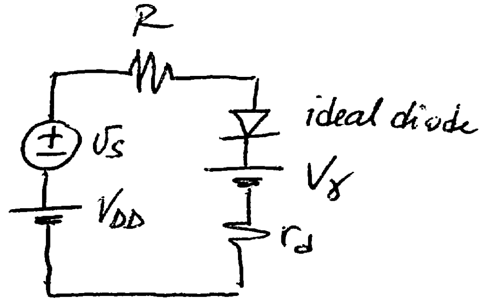
For small signals we can reasonably approximate the diode i-v curve with its tangent:



which is the piece-wise linear model that we already introduced.

The circuit we have analyzed is sort of problematic when we want to set I_{DQ} !!!
Pratically that's what we do:





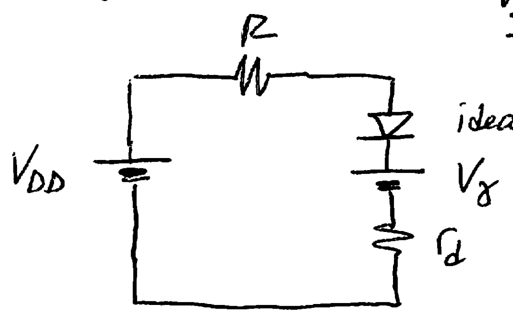
$$V_{DD} + v_S = R \cdot i_D + v_D$$

$$V_{DD} + v_S = R \cdot i_D + V_\gamma + r_d\, i_D$$

Superposition:

① only dc

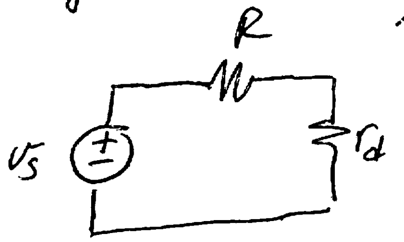
$\underline{V_{DD} > V_\gamma}$!!!



$$V_{DD} = I_D R + \underbrace{V_\gamma + r_d I_D}_{V_D}$$

② only ac

Basically for the small signals we model the diode with its incremental resistance!



$$v_S = i_d \cdot (R + r_d)$$

①+② :

$$V_{DD} + v_S = R\,(\underbrace{I_D + i_d}_{i_D}) + V_\gamma + r_d\,(\underbrace{I_D + i_d}_{i_D})$$

So this approach where we separated dc world and ac world (for small signals) makes sense !!!

DIODE HIGH FREQUENCY MODEL

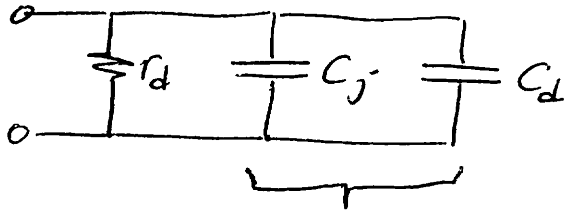
So far we have totally neglected the diode charge storage effects.

↓

This is reasonable only when the small signals we apply have a frequency that is relatively small

↓

If the frequency goes up we have to account for C_j and C_d



For low frequencies
C_j and C_d are open circuits
but for high frequencies
C_j and C_d becomes short circuits !!

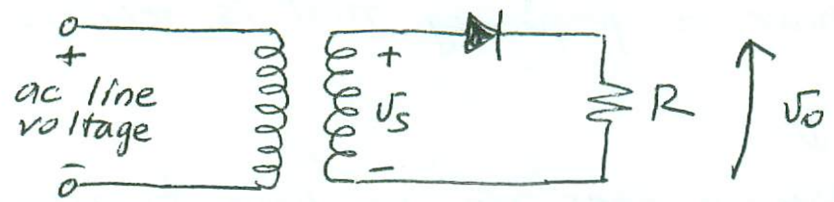
↓

It takes time to switch the diode ON-OFF and viceversa !!

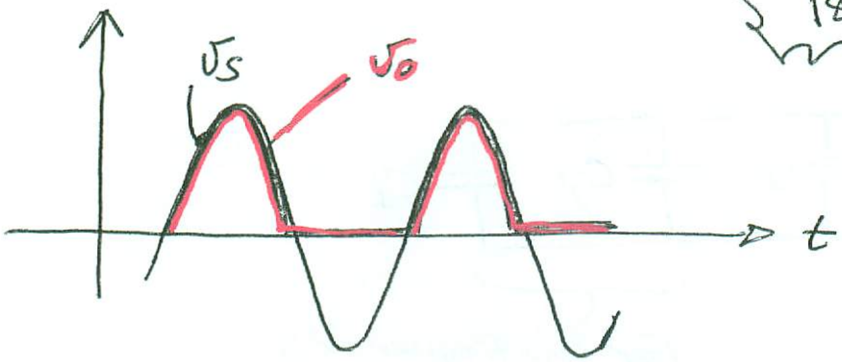
Elementary diode applications

- **Rectifiers**

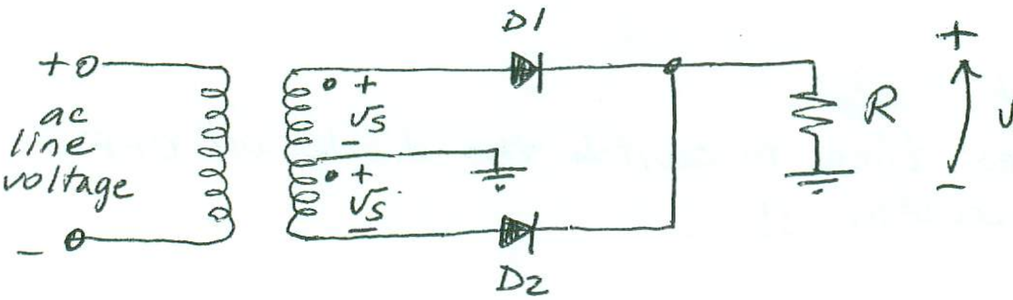
Half wave rectification



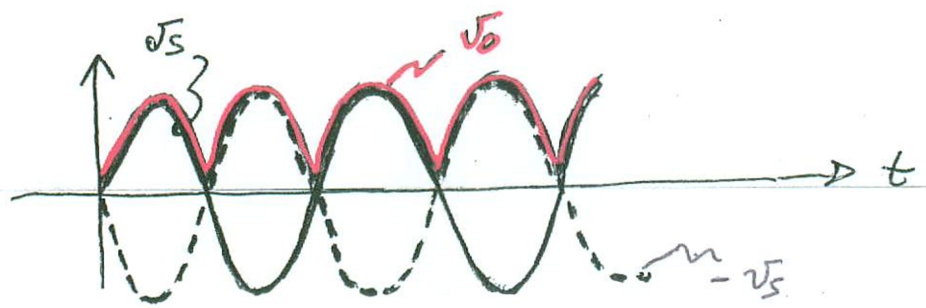
ERROR ON PAGE 181 OF TEXTBOOK



Full-wave Rectification



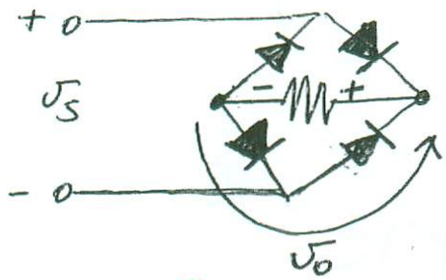
ERROR ON PAGE 183 OF TEXTBOOK



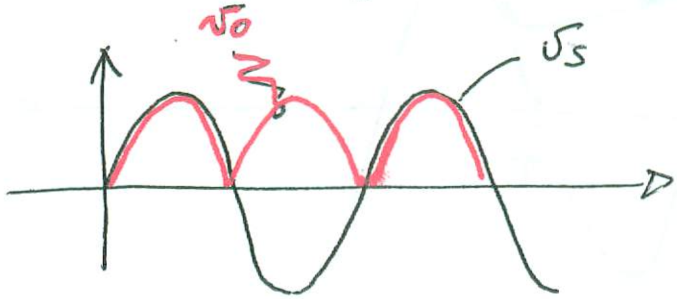
PIV of the diodes is about $2V_s$!!!

peak inverse voltage

bridge rectifier (Grätz)



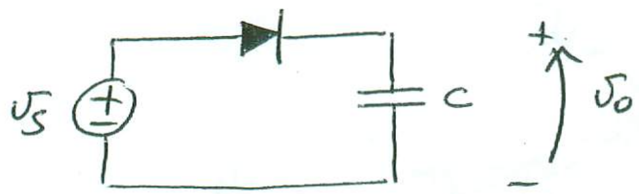
ERROR ON PAGE 184 OF TEXTBOOK

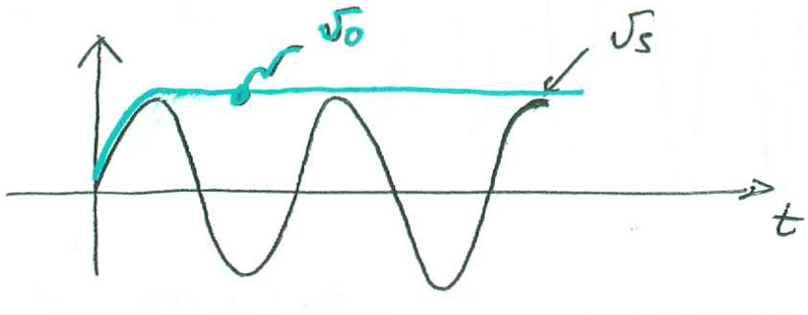


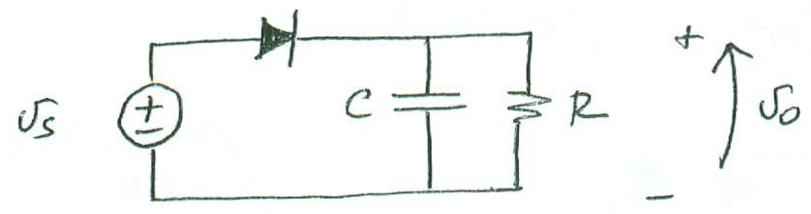
PIV of the diodes is about V_S !!

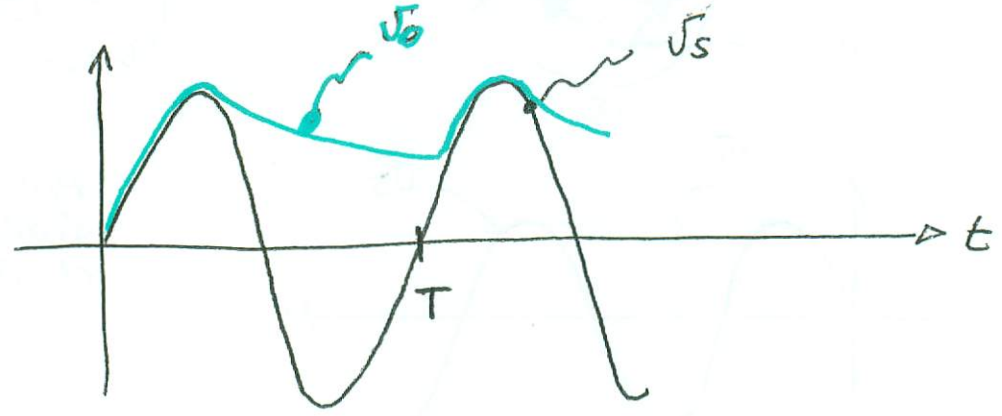
Peak Rectifier – (Rectifier with capacitive filter)

" peak detector (can be used to reconstruct the envelope of an AM signal)



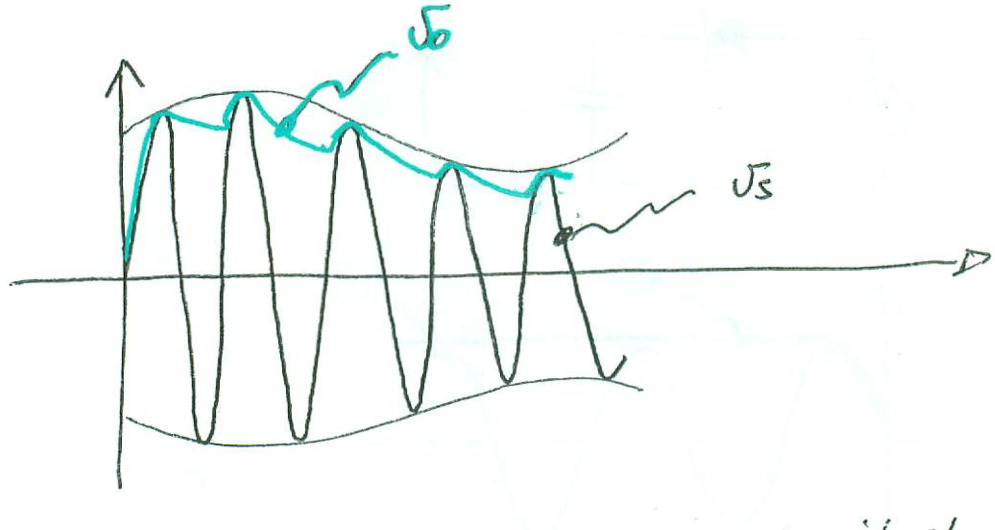






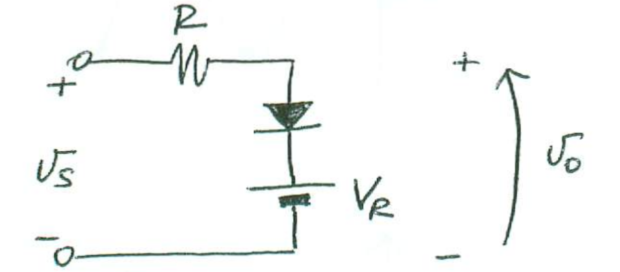
we have to choose $\rightarrow RC \gg T$

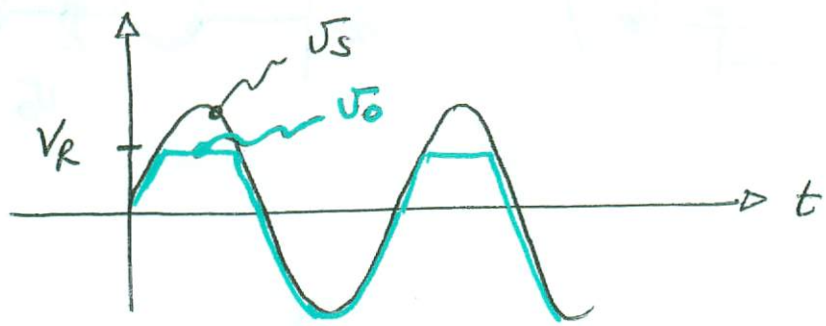
during the discharge $\rightarrow v_o = V_p \, e^{-t/RC}$

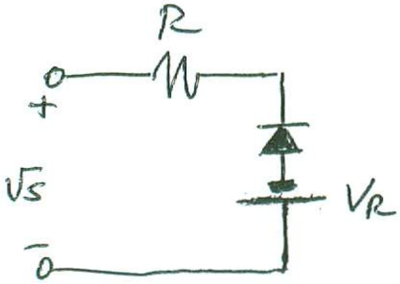


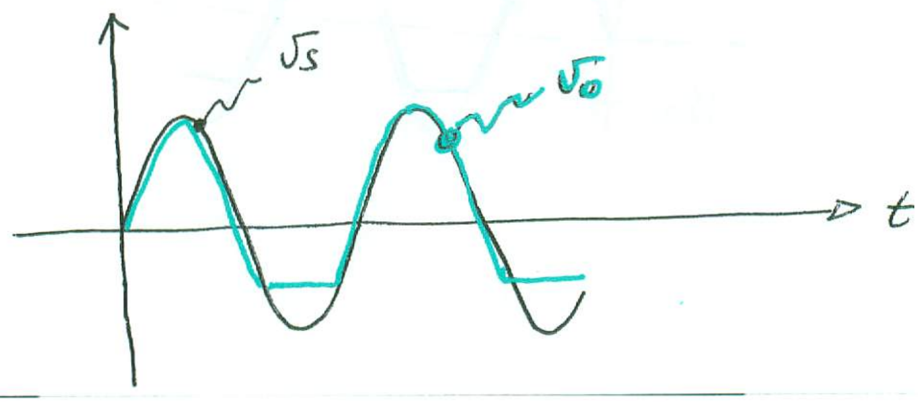
If I feed an AM signal to this circuit it will detect its envelope

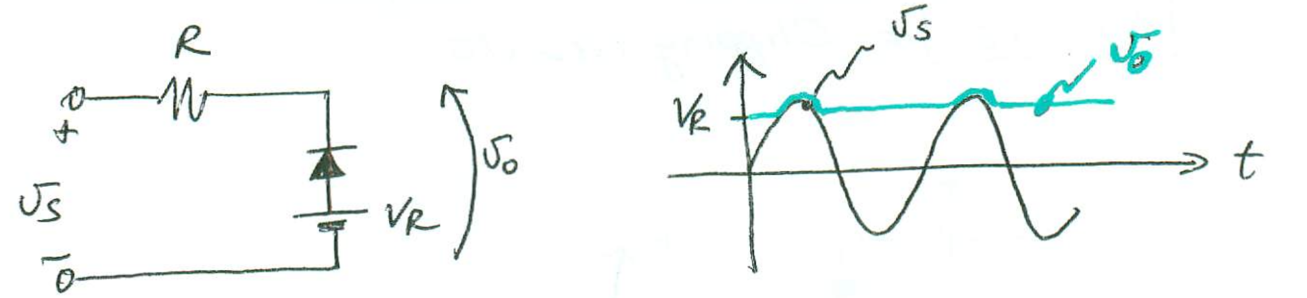
$\boxed{\text{Limiters}}$ = Clipping circuits

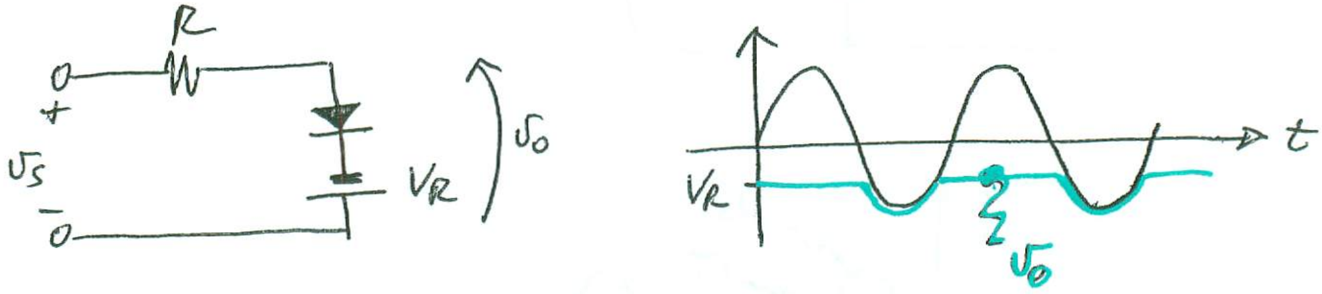


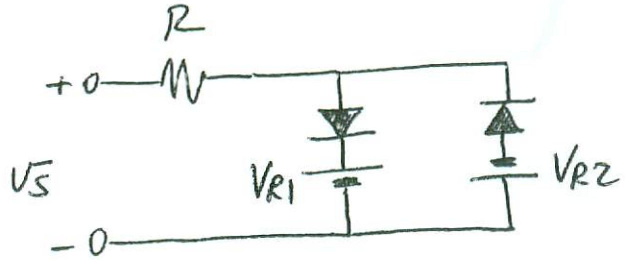


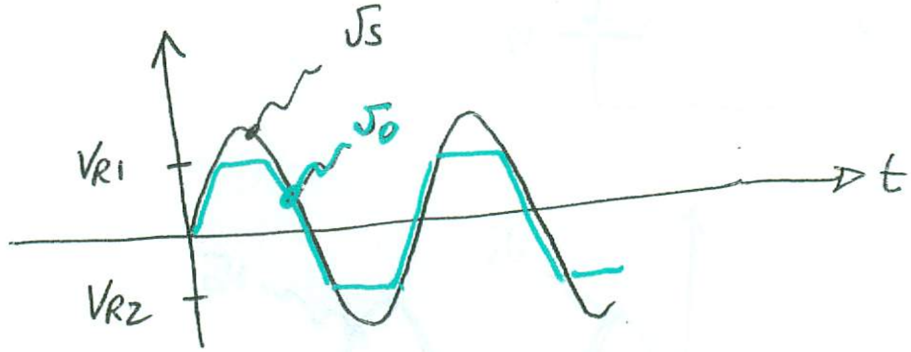





R
+
−
vs
vo
VR
t

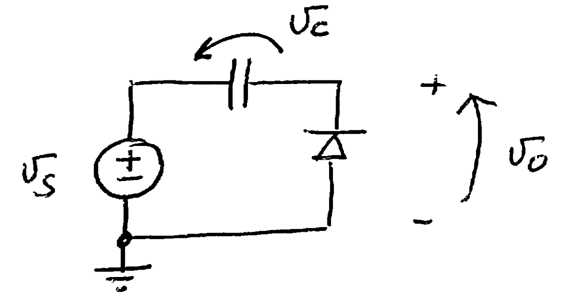

R
+
−
vs
vo
VR
t


R
+
−
vs
VR1
VR2


vs
vo
VR1
VR2
t

CLAMPING CIRCUITS

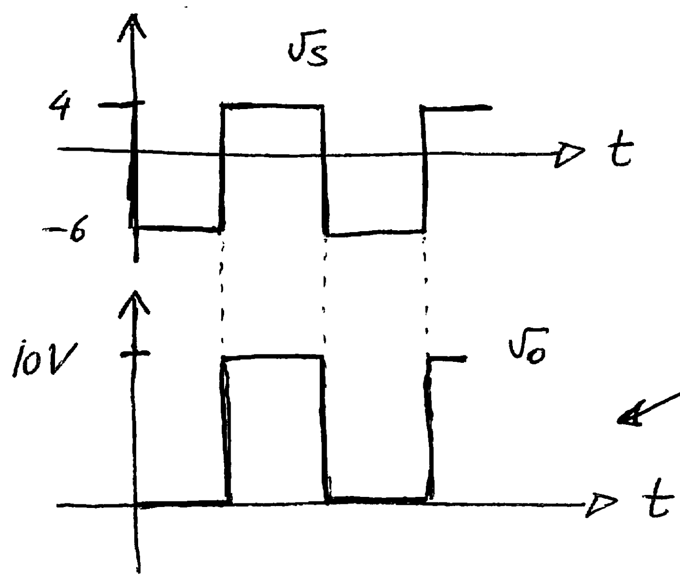
Clamping is a shifting operation of a waveform, where the amount of shift depends upon the actual waveform.



$$i_c = C\frac{dv_c}{dt}$$

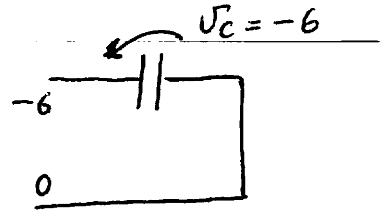
$$v_o = v_s - v_c$$

This is a though circuit to analyze

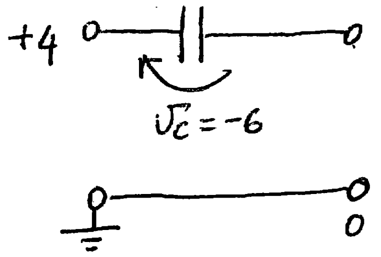


DC RESTORATION !!

At the beginning the diode is a short and the capacitance will charge to a voltage v_c equal to the magnitude of the most negative peak ($v_c = -6V$)

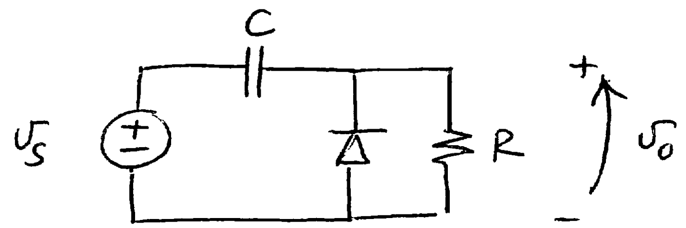


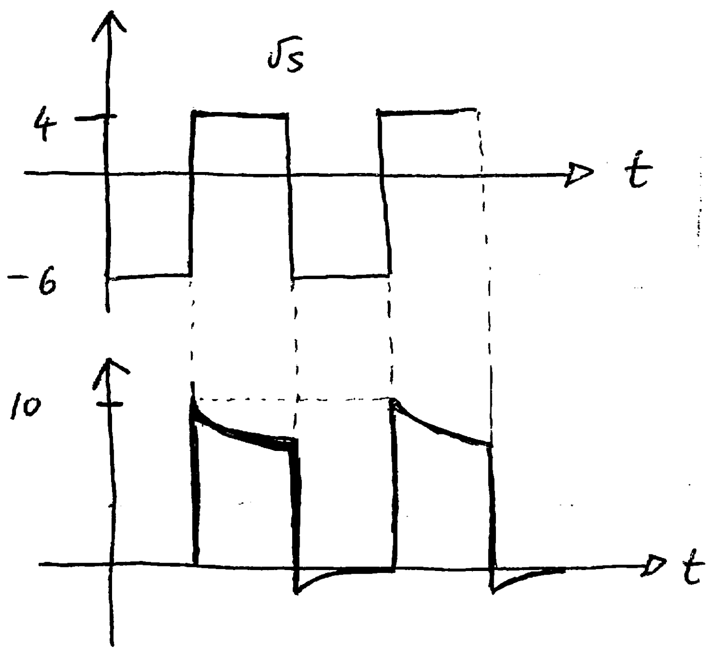
Subsequently the diode open and the capacitor retains its voltage indefinitely



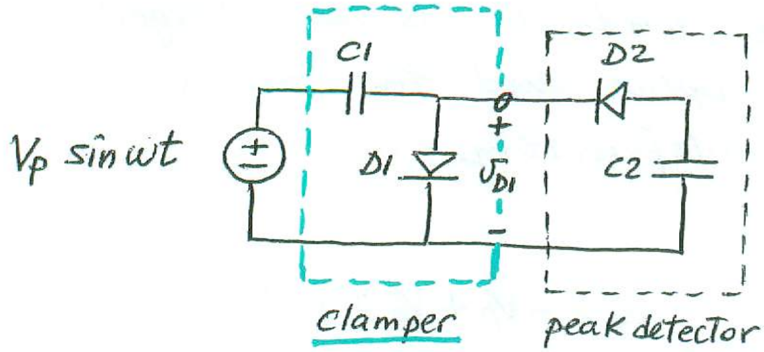
$v_o = v_i - v_c = 4 - (-6) = +10$

More pratically there will be a load resistance connected across the diode in a clamping circuit

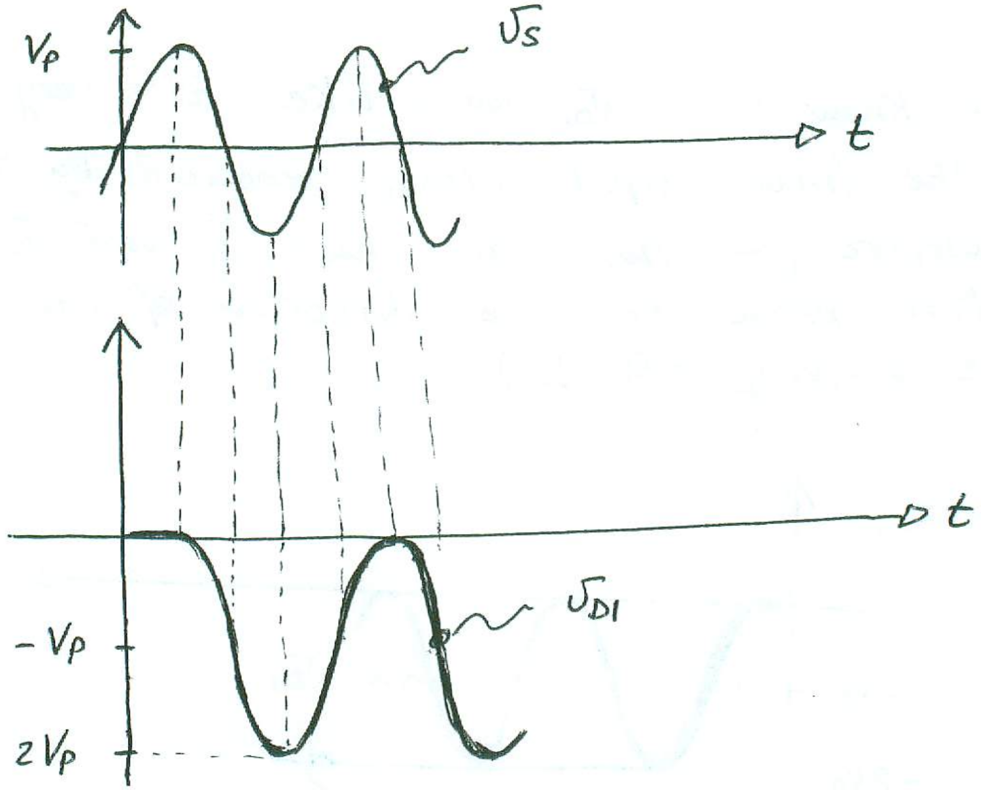




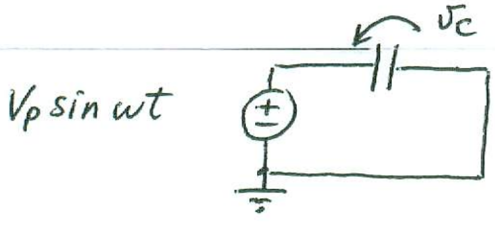
VOLTAGE DOUBLER



This is another tough circuit to analyze

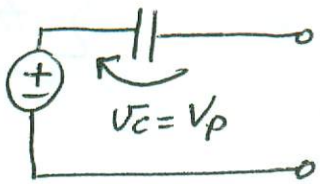


at the beginning the diode is a short and the capacitance will charge to a voltage v_C equal to V_p.



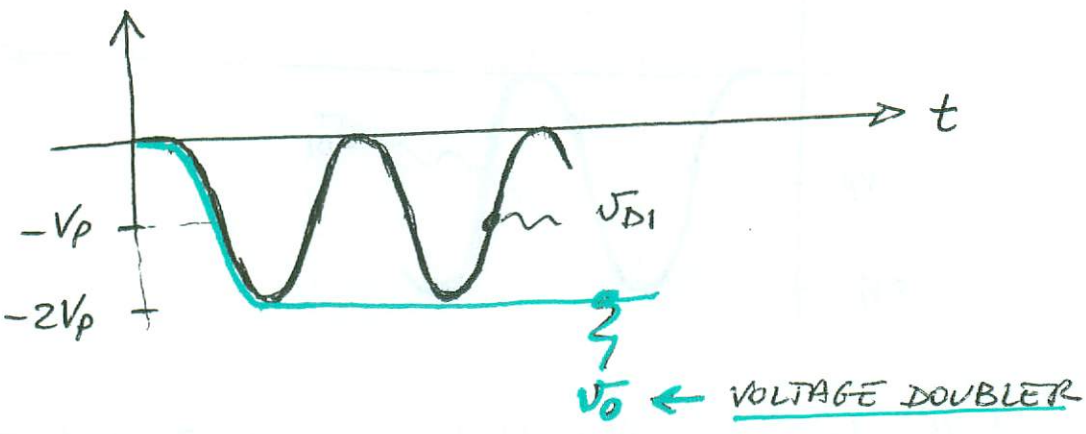
Subsequently after $\frac{\pi}{2}$ when the sine wave has reached is peak value and C_1 is now charged at $v_C = V_p$ the diode open and the capacitor retains its voltage indefinitely

$V_p \sin wt$



$$v_o = -V_p + V_p \sin wt$$

once we know how v_{D1} looks like it's very easy to get the final output voltage produced by the peak detector ($\rightarrow$ note that here I want to detect a negative peak so the direction of the diode D2 must be properly set !!)

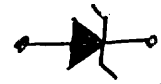


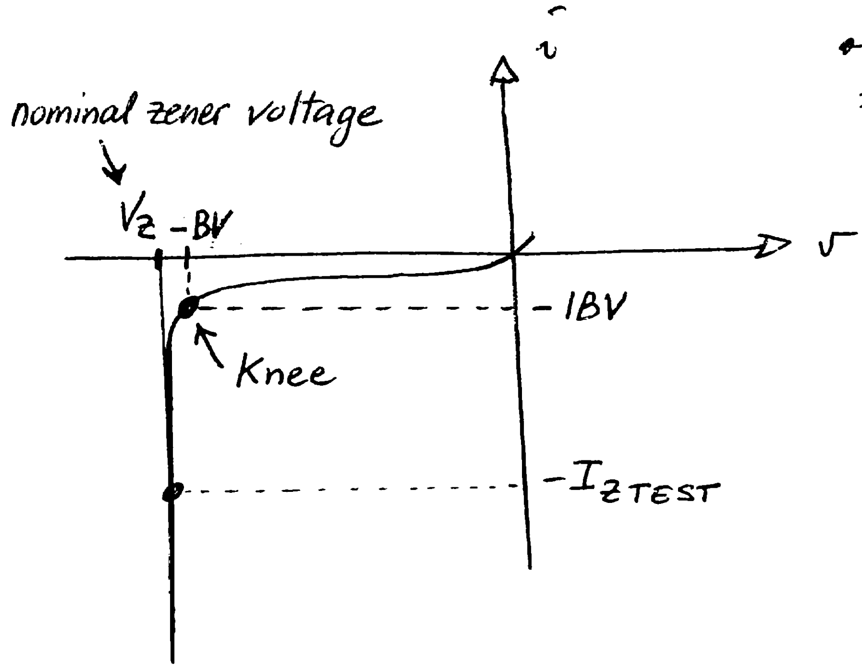
ZENER DIODES

A zener diode is a device where the doping is performed in such a way as to make the i-v characteristic in breakdown region very steep.
If the reverse voltage exceeds the breakdown voltage the diode normally will not be destroyed.
(the manufacturer specifies the maximum power that the device can safely dissipate → sometimes they give directly the max current that the device can safely tolerate)

↓

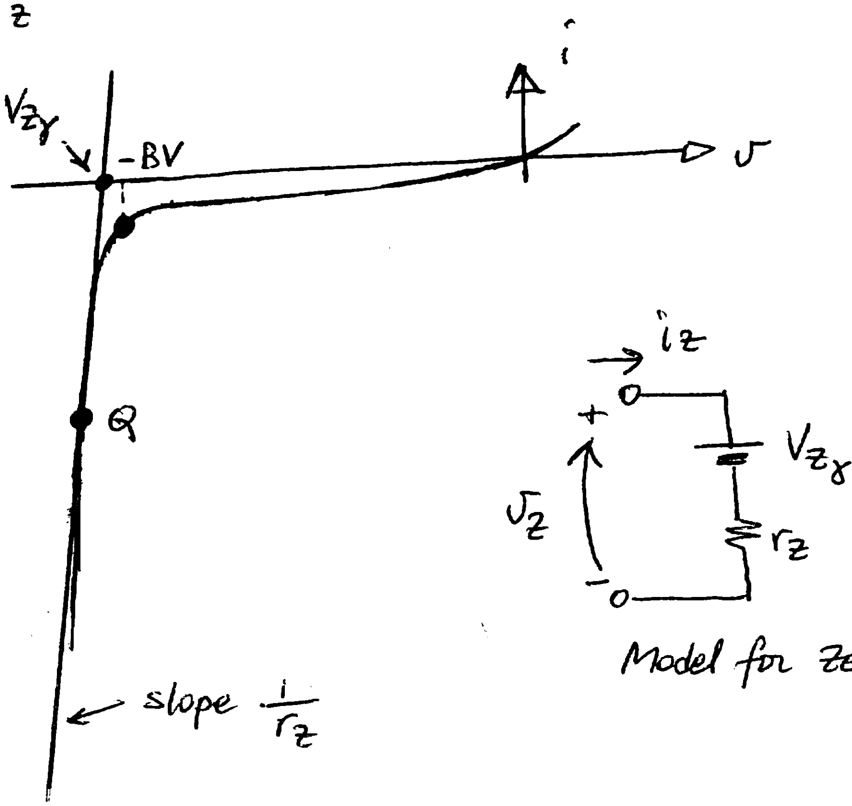
the almost constant voltage drop that the diode exhibits in the breakdown region suggest that a natural application could be in the design of voltage regulators.





~~The manufacturer usually specifies~~ the "nominal" voltage across the zener V_Z at a specified test current I_{ZTEST}.

Though the $i-v$ curve is very steep in the breakdown region it is not vertical $\rightarrow$ it has a certain slope $\frac{1}{r_Z}$



Model for ZENER diode

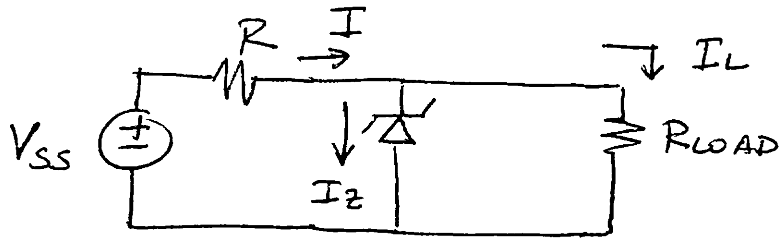
pratically it is often assumed that $V_{Z\gamma} \approx \underbrace{-BV}_{\text{knee voltage}}$

The maximum reverse current I_{Zmax} that the zener can withstand is dependent upon the design and construction of the diode.

The leakage current (I_{Zmin}) at the knee of the caracteristic curve is sometime assumed 10% of the I_{Zmax} ($\rightarrow$ I_{Zmin} is what we also called IBV)

in lack of more accurate data !!!

Example



It's given a supply voltage of nominally 10V but that can vary by ± 1V.

We want to design a voltage regulator able to provide a constant voltage of 6.8 V to a load R_L.

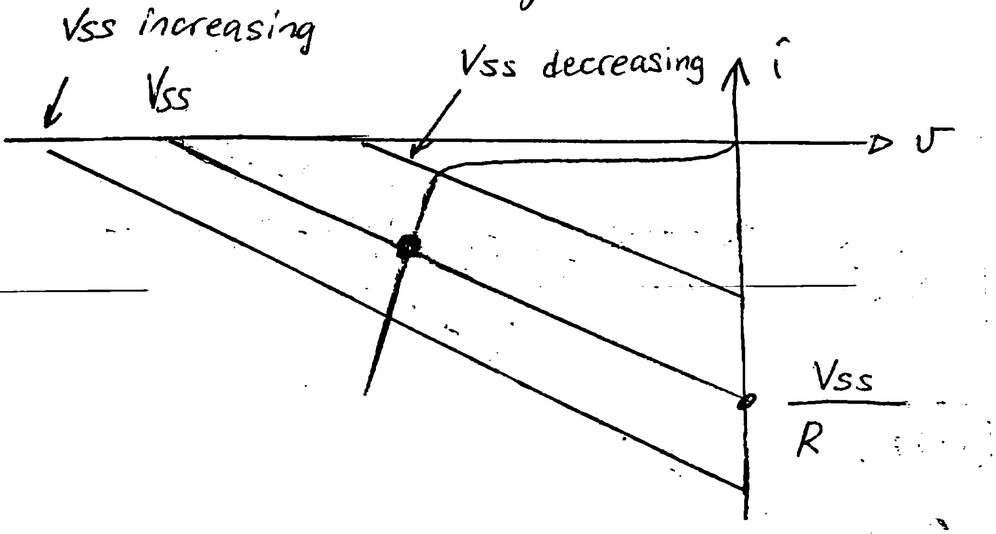
The load can vary from 0.5 KΩ to 2KΩ.

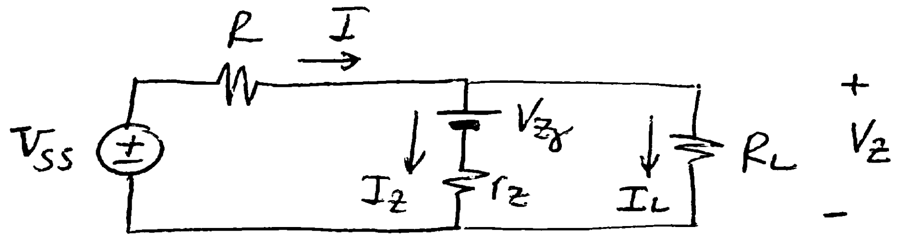
We have a zener specified to have $V_Z = 6.8$ at $I_Z = 20$ mA. r_Z is about 5 Ω and the max current that the zener can withstand is about 60 mA

(ZENER 1N754)

↓

in order to properly design this circuit the resistance R must be such that the diode stays in the constant voltage region over the entire range of input voltages and load impedance !!





$$\begin{cases} V_{SS} - V_Z = R \cdot I \\ I = I_Z + I_L \end{cases} \longrightarrow I = I_Z + \frac{V_Z}{R_L}$$

In order to assure that the diode remains in the constant voltage region, there are two extreme conditions to consider:

- The highest current through the diode occurs when the source voltage is maximum and R_L is maximum ($\rightarrow$ the worst thing that can happen is that the load get unconnected because then all current supplied I goes through the diode)

$$I_{Z\,max} = \underbrace{\frac{V_{SS\,max} - V_Z}{R}}_{= I} - I_{L\,min} =$$

$$= \frac{V_{SS\,max} - V_Z}{R} - \frac{V_Z}{R_{L\,max}} \xrightarrow[R_L \to \infty]{}$$

$$\simeq \frac{V_{SS\,max} - V_Z}{R}$$

- The lowest current through the diode occurs when the source voltage is minimum and R_L is minimum ($\rightarrow$ most of current supplied I is drawn by R_L)

$$I_{Z_{min}} = \frac{V_{SSmin} - V_Z}{R} - I_{L_{max}} =$$

$$= \frac{V_{SSmin} - V_Z}{R} - \frac{V_Z}{R_{Lmin}}$$

Let's try now to select R based on the nominal values ($\rightarrow$ intermediate case) but assuming that there is no load ($\rightarrow$ all current provided by the power supply will flow through the zener $\rightarrow$ we are over-pessimistic $\rightarrow$ just to be be on the safe side)

$$I_{Z_{nom}} = \frac{V_{SS} - V_{Znom}}{R} \rightarrow R = \frac{V_{SS} - V_{Znom}}{I_{nom}}$$

$$\left.\begin{aligned} I_{Znom} &= 20\ mA \\ V_{Znom} &= 6.8\ V \\ V_{SS} &= 10\ V \end{aligned}\right\} \rightarrow R \simeq 160\ \Omega$$

Let's now verify if this choice of R will satisfy the 2 extreme situations!!

1. $$I_{Z_{max}} \leq \frac{V_{SSmax} - V_Z}{R} \simeq 27\ mA$$

The zener can stand up to 60 mA so that's fine!

② $$I_{Z\,min} \geq \frac{V_{SS\,min} - V_Z}{R} - \frac{V_Z}{R_{L\,min}} =$$

$$= \frac{9-6.8}{160} - \frac{6.8}{500} \simeq 0.2\ mA$$

HERE we have a problem !!
The current at the knee is about 6mA (→ 10% of 60 mA) so we are not in the constant voltage region

we need to lower R !!

In order to remain in the constant voltage region we need to choose R in such a way that:

$$I_{Z\,min} + I_{L\,max} = \frac{V_{SS\,min} - V_Z}{R}$$

$$R \leq \frac{V_{SS\,min} - V_Z}{I_{Z\,min} + I_{L\,max}} = \frac{9-6.8}{6\times 10^{-3} + \frac{6.8}{500}} = \simeq 112\ \Omega$$

I need more current, so more current will flow through the zener

Let's than take $R = 100\ \Omega$ and cross-check what happen:

② $I_{Z\,min} = \frac{9-6.8}{100} - \frac{6.8}{500} \simeq 8.4\ mA$

① $I_{Z\,max} = \frac{10-6.8}{100} \simeq 42\ mA$

→ ok !!

5.32

In order to evaluate how good is the voltage regulator we designed we can measure the total voltage swing divided the nominal voltage (provided to the load)

$$\text{PERCENT REGULATION} = \frac{\Delta V_o}{V_{o_{nom}}} \times 100$$

The goal was to provide a nominal voltage of 6.8 V to the load

$$\Delta V_o = V_{o_{max}} - V_{o_{min}} = r_z \cdot \Delta I_z$$

$$\Delta I_z = I_{z_{max}} - I_{z_{min}}$$

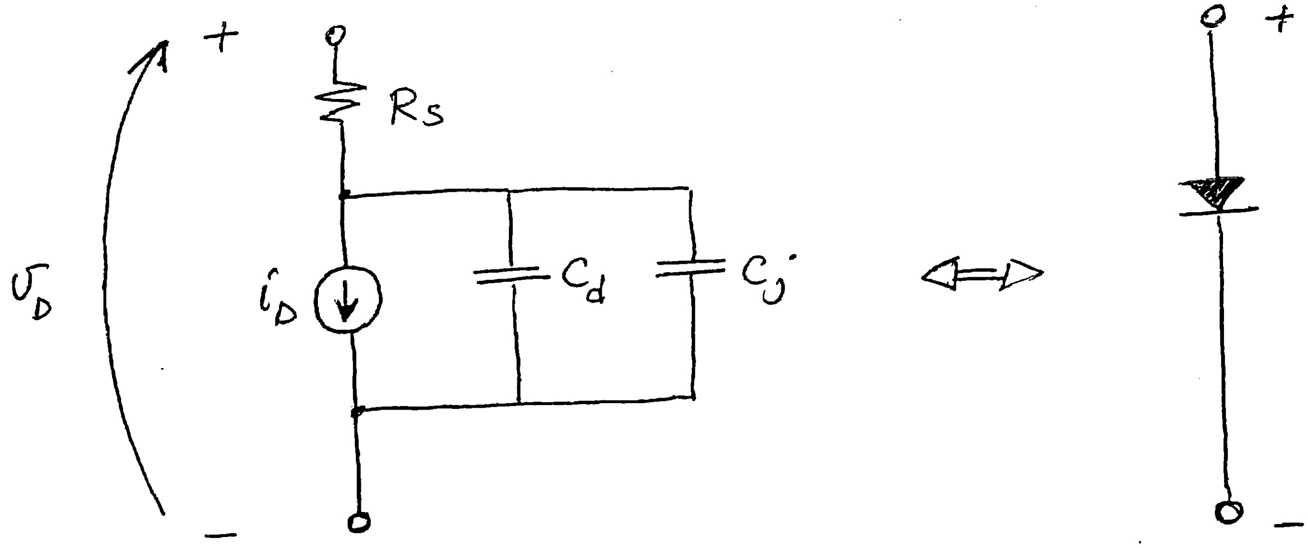
$$\left.\begin{aligned} I_{z_{max}} &= 42\ mA \\ I_{z_{min}} &= 8.4\ mA \end{aligned}\right\} \rightarrow \Delta I_z \simeq 33.6\ mA$$

$$\Delta V_o = \Delta I_z \cdot r_z \simeq 33.6\ mA \cdot 5\Omega \simeq 168\ mV$$

$$\%\ \text{REGULATION} = \frac{\Delta V_o}{V_{o_{nom}}} \cdot 100 = \frac{168\ mV}{6.8} \cdot 100 \simeq 2.6\%$$

NOT TOO BAD !!

DIODE SPICE MODEL



$$i_D = I_S \left(e^{v_D/\eta V_T} - 1 \right)$$

$$C_d = \frac{\tau_T}{V_T} \underbrace{I_S e^{v_D/\eta V_T}}_{g_d}$$

$$C_j = \frac{C_{jo}}{\left(1 - \frac{v_D}{\phi_B}\right)^m}$$

$$C_{jo} = \frac{A}{\sqrt{\frac{2}{q\varepsilon_s}\left(\frac{N_A + N_D}{N_A \cdot N_D}\right)\phi_B}}$$

FROM: SEDRA, SMITH

Table 3.3 SPICE DIODE MODEL PARAMETERS (SOME OF THEM)

Model Parameter	Symbol	SPICE Name	Units	Default Value
Saturation current	I_S	IS	A	1×10^{-14}
Emission coefficient	n	N	—	1
Ohmic resistance	R_S	RS	Ω	0
Built-in voltage	Φ_B	VJ	V	1
Zero-bias junction capacitance	C_{j0}	CJ0	F	0
Grading coefficient	m	M	—	0.5
Transit time	τ_T	TT	s	0
Breakdown voltage (KNEE)	V_{BR}	BV	V	∞
Reverse current at V_{BR} (KNEE)	I_{BR}	IBV	A	1×10^{-10}

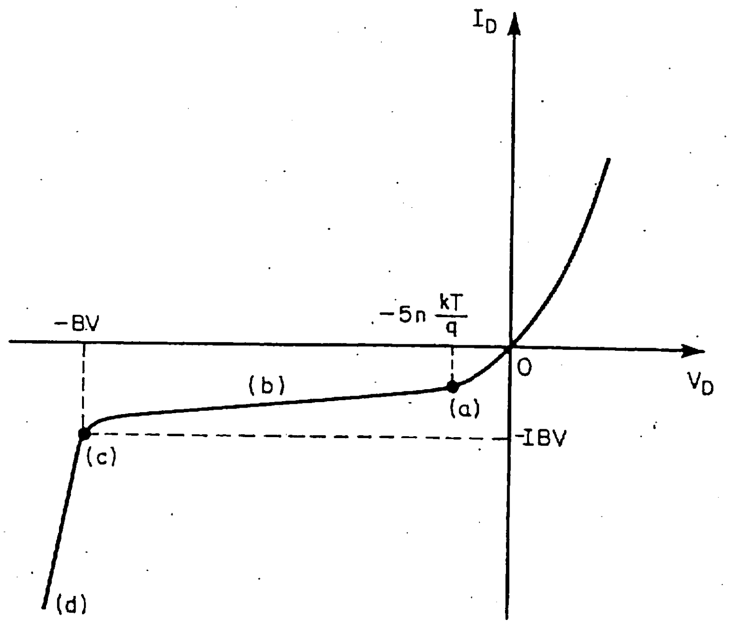


Figure 1-15 Reverse characteristic of the real diode.

From: P. ANTOGNETTI, G. MASSOBRIO
Semiconductor device modeling with SPICE
McGraw-Hill, New York, 1988

HOW TO BUILD A SPICE MODEL FOR A ZENER

Here D_1 is an ideal diode that can be implemented in SPICE by using a very small value for n (say $n = 0.01$), and D_2 is a regular diode model for the forward direction of the zener (for most applications the parameters of D_2 are of little consequence).

From: Sedra, Smith

since we do not usually use a ZENER in the forward direction we put very little attention in how D2 is modeled (→ default values are fine !!)

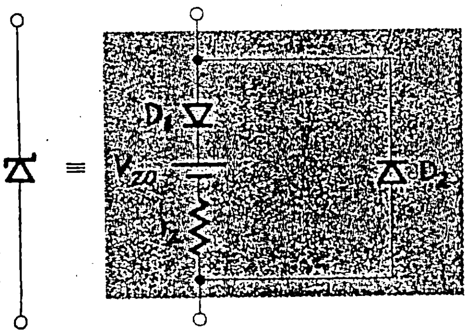
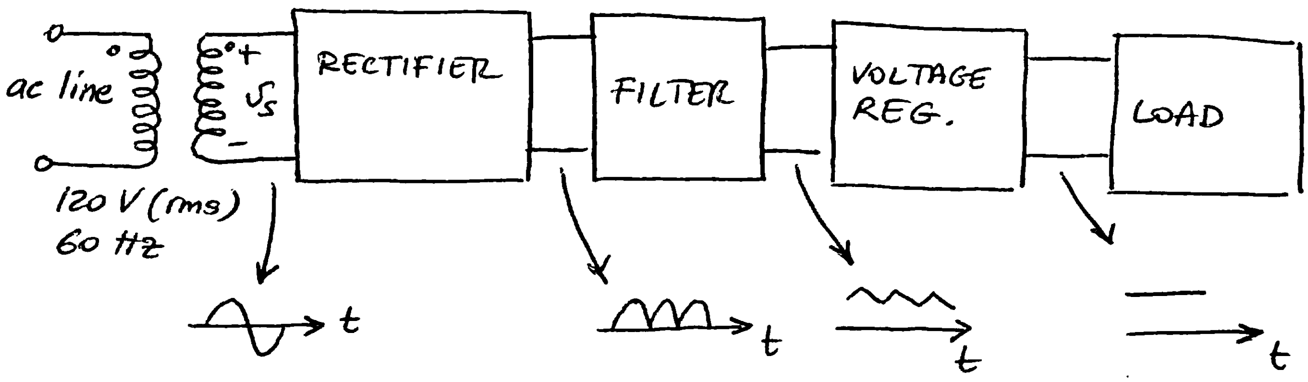


Fig. 3.51 Model for the zener diode. This model can be used in SPICE by defining the zener as a subcircuit. Diode D_1 is ideal and can be approximated in SPICE by using $n = 0.01$.

DIODE DESIGN EXAMPLES

- DC POWER SUPPLY DESIGN



$v_s \rightarrow 120\left(\frac{N_2}{N_1}\right)$ volt rms

The transformer besides providing the desired voltage transformation so that we have the desidered sine wave amplitude, provides electrical isolation. ($\rightarrow$ If I have a short at the primary the primary winding will burn, but no damages to the electronic equipment attached at the secondary will occurr)

We want to design a power supply that provide a dc voltage of nominally 5V and that is able to supply up to 25 mA to the Load.
The load can range from $R_L = 200 \div 500\,\Omega$
We have available a zener of $V_{Z\,nom} = 5.1\,V$ at $I_{Z\,nom} = 20\,mA$ and $r_z = 10\,\Omega$.
The maximum current that the zener can stand is $I_{Zmax} = 50\,mA$ ($\rightarrow$ the current through the zener should be at least 5 mA to be sure that we are in the constant voltage breakdown region $\rightarrow I_{Zmin} = 10\%\ I_{Zmax} = 5\,mA$)

Let's start our design.

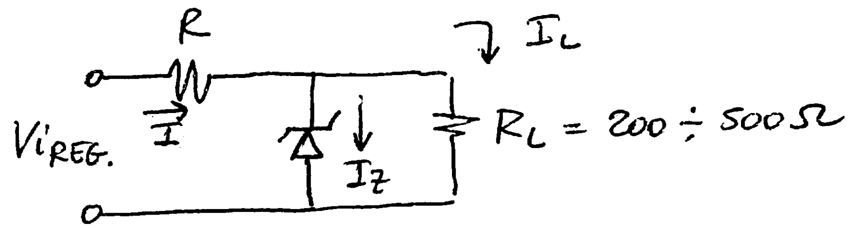
↓

The goal is to achieve the specified requirements on the load → so let's focus on the load and proceed with the design "backward" block by block from the load to the ac line

↓

$V_L = 5V$ dc.

$I_{L_{max}} = 25$ mA



First trivial observation is that if I want a constant voltage of 5V on the load I need a zener which has a nominal zener voltage "close" to 5V → we go through data sheets and we have been lucky → we found a zener which has:

$V_{Z_{nom}} = 5.1V$

$I_{Z_{nom}} = 20$ mA

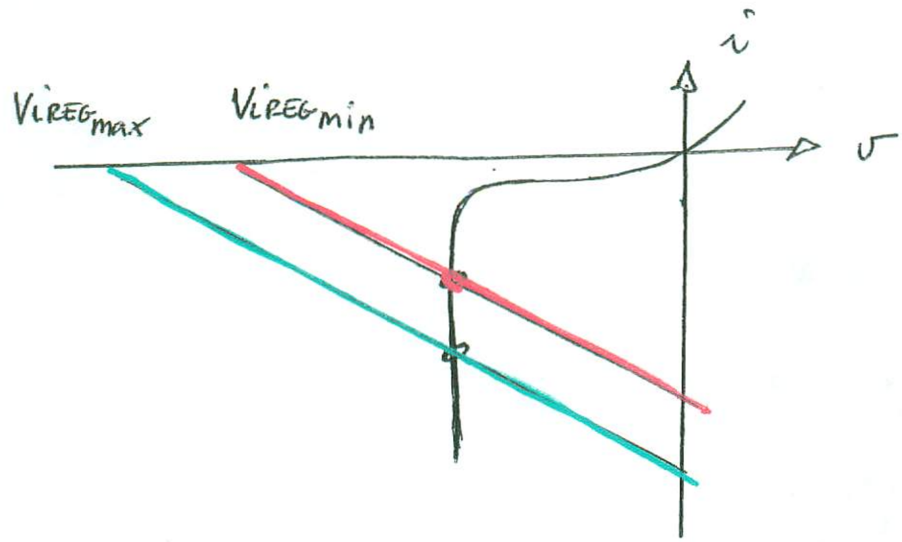
$r_Z = 10\Omega$

$(I_{Z_{max}} = 50$ mA, $I_{Z_{min}} = 5$ mA$)$

↓

LOAD LINE CONCEPT

If I want to make sure that the zener is always operating in the constant breakdown voltage region the Vi_{REG} must always be relatively bigger than the $V_{Z_{nom}}$ → at least a couple of times



Besides V_{iREG} is "fluctuating" → (if it were not I would have not spend my time trying to regulate it !!!)

How much is "fluctuating" I do not know, but it's up to me really !!! It depends on how good I am in filtering after the rectification.
At the moment I can't say for sure, but I am reasonably confident that it shouldn't be that big deal to mantain the fluctuations within a couple of volts → We'll cross check it later if we're able to satisfy this assumption !!!

↓

$$V_{iREG\,min} = 10\,V$$

$$V_{iREG\,max} - V_{iREG\,MIN} = 2\,V \leftarrow \underline{\text{RIPPLE}} = \Delta V$$

$$V_{i_{REG}} = R \cdot I + V_Z$$

$$I = I_L + I_Z$$

$\Downarrow$ Let's choose R !

①

$$V_{i_{REG\,min}} = R \cdot I_{min} + V_Z$$

$$I_{min} = \underbrace{I_L}_{\substack{\shortparallel \\ 25\,mA}} + \underbrace{I_{Z\,min}}_{\substack{\shortparallel \\ 5\,mA}}$$

we always want to be able to supply up to 25 mA to the load

$$\downarrow$$

$$10 = R \cdot 30\,mA + 5.1 \longrightarrow R \cong 163\,\Omega$$

②

$$V_{i_{REG\,max}} = R \cdot I_{max} + V_Z$$

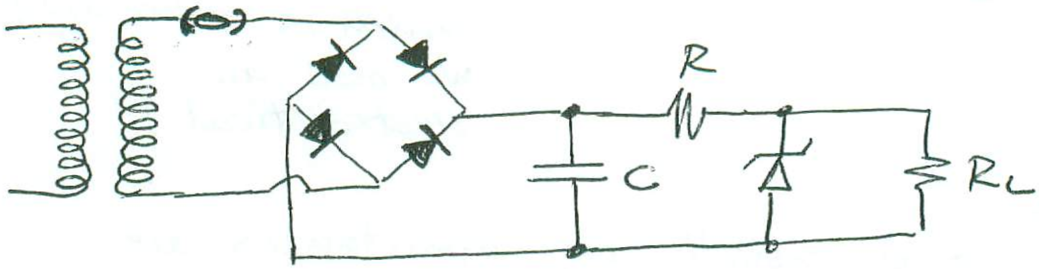
$$I_{max} = \frac{V_{i_{REG\,max}} - V_Z}{R} = \frac{12 - 5.1}{163} \simeq 42\,mA$$

From zener point of view the worst thing that can happen is that the load get unconnected $\longrightarrow$ then all 42 mA go through the zener $\longrightarrow$ the zener can stand up to 50 mA so we are safe !!!

In normal conditions only $42 - 25 = 17$ mA goes through the zener !

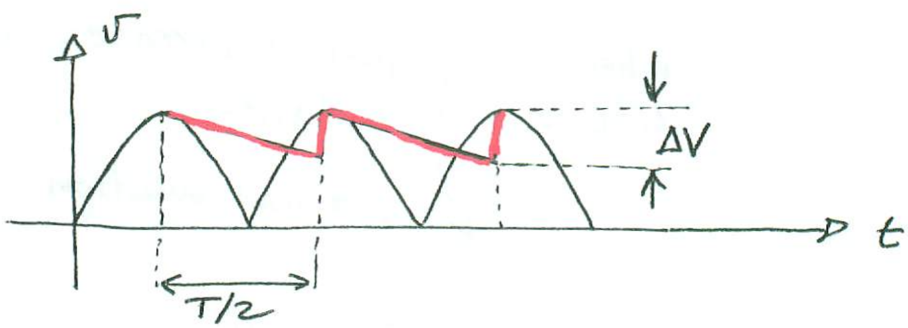
$$P_{R_{max}} = \frac{V_R^2}{R} = \frac{(12-5.1)^2}{163} = 292\ mW$$

$$P_{Z_{max}} = V_Z \cdot I_{Z_{max}} \cong 5.1V \cdot 42\ mA \simeq 214\ mW$$



Let's now choose C so that we can be sure that the ripple stays below 2V ($\Delta V \leq 2V$)

$\Delta V = 2V$



$$I = \frac{\Delta V}{\Delta t} \cdot C \longrightarrow C = \frac{\Delta t}{\Delta V} \cdot I = \frac{T/2}{\Delta V} \cdot I \longrightarrow$$

pessimistic assumption that the discharge last T/2

$$C = \frac{T}{2 \cdot \Delta V} \cdot I \cong \frac{T}{2 \cdot \Delta V} \cdot I_{max}$$

if I take C big the discharge is slow (which is what I want!!)

oversizing the capacitance C I am sure that the ripple is going to be definitely better !!!

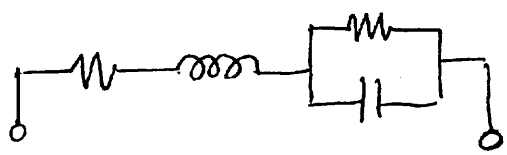
↓

$$C \simeq \frac{1/60}{2 \cdot 2} \cdot 42 \times 10^{-3} \simeq 200\,\mu F$$

↑

which is rather big !!!
we need an electrolitical !!

→ usually big capacitances are anything but capacitances

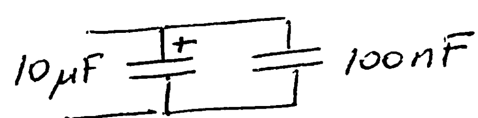


$$X = j\omega L - j\frac{1}{\omega C}$$

↓

when the frequency increases it behaves like an inductance

↓ pratical solution



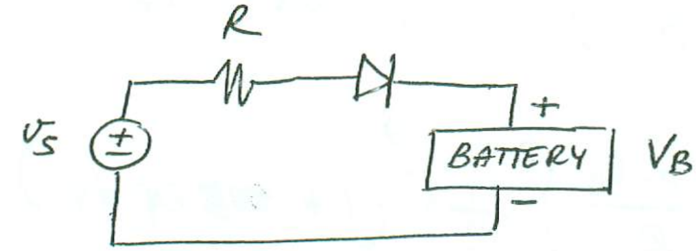
↑ at low frequency this keep the situation under control.

↖ at high frequency this keep the situation under control

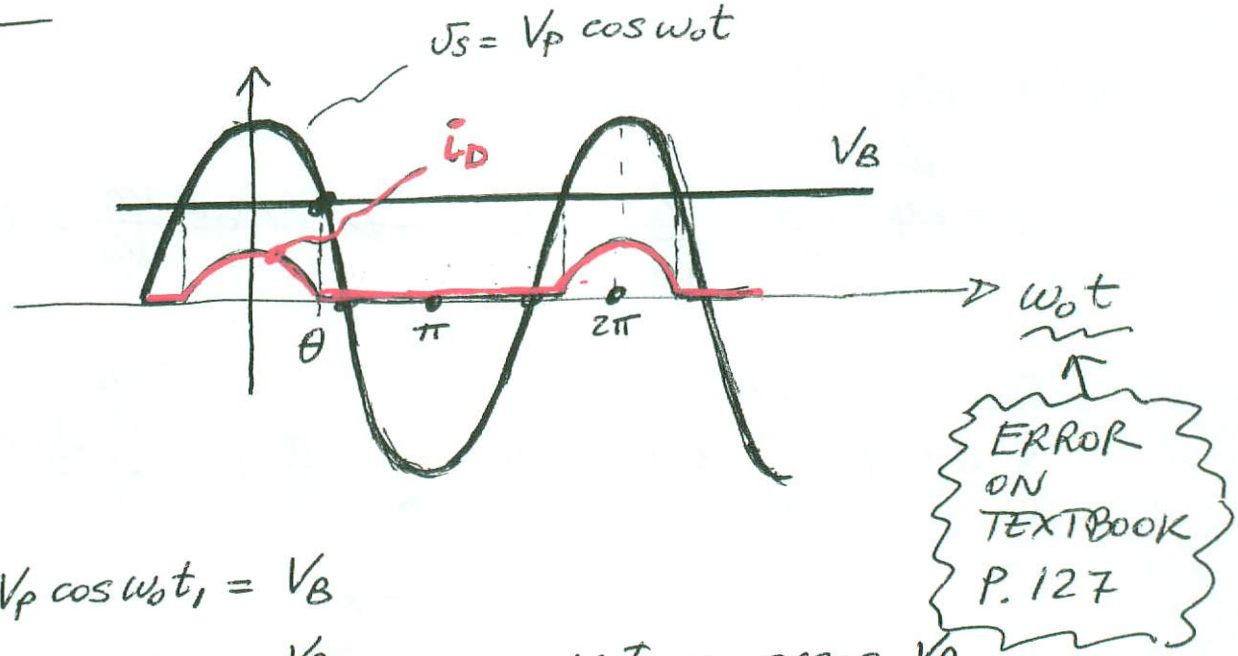
Anyway we are working at 60 Hz so there is no problem.

Regarding the transformer we need 12 V peak at the secondary so a ratio of $\frac{120 \cdot \sqrt{2}}{12} \approx 14$ is ok !!

- BATTERY CHARGER



We want to design a battery charger for a battery of 12 V dc that supplies 100mA dc



$$V_P \cos \omega_o t_1 = V_B$$

$$\cos \omega_o t_1 = \frac{V_B}{V_P} \longrightarrow \underbrace{\omega_o t_1}_{= \theta} = \arccos \frac{V_B}{V_P}$$

$$I_{DPEAK} = \frac{V_P - V_B}{R}$$

$\varphi = \omega_o t$

$$I_{Drms}^2 = \frac{2}{2\pi} \int_0^\theta I_{DPEAK}^2 \cdot \cos^2 \varphi \, d\varphi =$$

$$= \frac{1}{\pi} \int_0^\theta I_{DPEAK}^2 \cos^2 \varphi \, d\varphi$$

$$I_{Drms}^2 = \frac{1}{\pi}\left(\frac{V_P - V_B}{R}\right)^2 \cdot \int_0^{\theta} \cos^2\varphi \, d\varphi =$$

$$= \frac{1}{\pi}\left(\frac{V_P - V_B}{R}\right)^2 \cdot \frac{1}{2}\int_0^{\theta}(1 + \cos 2\varphi \, d\varphi)$$

We want:

$I_{Drms} = 100 \text{ mA}$ $\qquad I_{DPEAK}$

let's fix V_P:

$$V_P = 24 \rightarrow \frac{V_B}{V_P} = \frac{1}{2} \rightarrow \arccos\frac{V_B}{V_P} = \theta = 60^\circ = \frac{\pi}{3}$$

$$\int_0^{\pi/3} d\varphi + \int_0^{\pi/3} \cos 2\varphi \, d\varphi = \frac{\pi}{3} + \int_0^{2\pi/6} \frac{1}{2}\cos x \, dx =$$

$2\varphi = x$

$d\varphi = \frac{dx}{2}$

$$= \frac{\pi}{3} + \frac{1}{2}\Big[\sin x\Big]_0^{\pi/3} = \frac{\pi}{3} + \frac{1}{2}\left[\frac{\sqrt{3}}{2} - 0\right] =$$

$$= \frac{\pi}{3} + \frac{\sqrt{3}}{4}$$

$$100^2 \times 10^{-6} = \frac{1}{\pi}\left(\frac{\pi}{3} + \frac{\sqrt{3}}{4}\right) \cdot \frac{1}{2}\left(\frac{12^2}{R^2}\right) \rightarrow$$

$$10^{-2} \cong 0.24 \cdot \frac{144}{R^2} \rightarrow R \simeq 59 \ \Omega$$

$$I_{D\,PEAK} = \frac{24-12}{55} = \simeq 218\ mA$$

PIV on the diode

↓

$12 + 24 = 36$

SPECIAL TYPES OF DIODES

- Schottky barrier diodes

The junction formed by a metal and a doped semiconductor can be either rectifying or ohmic. (because of the difference in carrier concentrations in the two materials, a potential barrier exists → in ohmic contacts care is taken to eliminate the effect of the barrier)

↓

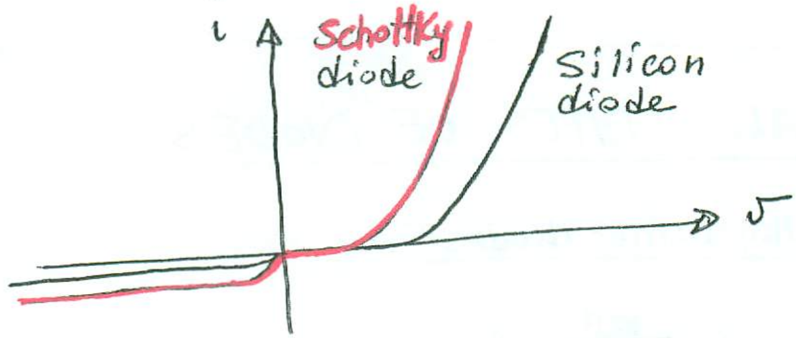
Schottky diodes are formed bringing together a metal (typically aluminium or platinum) and an n-type semiconductor material (silicon).

↓

in schottky diodes the current is conducted by majority carriers (→ this means that there is no charge-storage effect due to the minority carriers → a schottky diode can be switched on-off faster)

The reason why the current is conducted by majority carriers is that in the metal there are available numerous electrons!

The forward voltage drop of schottky diodes is smaller $0.3 \div 0.5\,V$, and the reverse saturation current is higher



- VARACTOR

A varactor is a diode specifically manufactured to be used as voltage variable capacitor.

A biased pn junction exhibits a certain capacitance that is function of the applied voltage

$$C_j = \frac{C_{j0}}{\left(1 - \frac{v_D}{\phi_B}\right)^m}$$

$$C_d = \frac{\tau_T}{V_T} e^{v_D/\eta V_T}$$

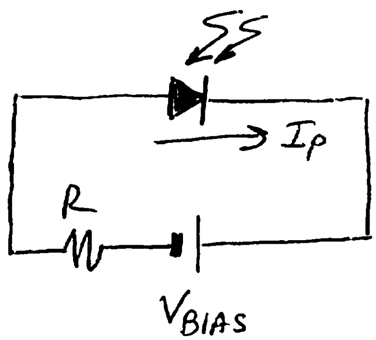
usually they are used with a reverse bias !!!
→ no current flowing through the device
and I can get a bigger ΔC enlarging the depletion region (→ rather than trying to shrink it applying a forward bias).

- Photodiodes

If a pn junction is illuminated the photons impacting the junction cause covalent bonds to break, and thus electron-hole pairs are generated (→ which induce current)

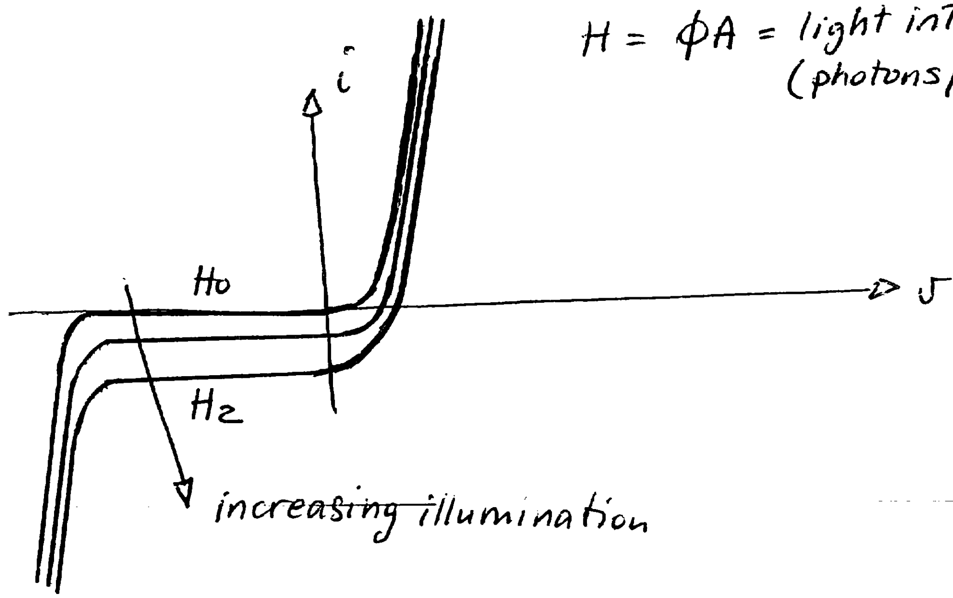
Photodiodes are usually fabricated using GaAs.

A photodiode converts light to electrical energy (→ using the photodiode forward bias we have a solar-cell → photodiodes for solar cells are usually fabricated from rather cheap silicon)



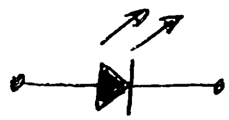
$I_p = \eta q H$

η = quantum efficiency
q = electron charge
ϕ = photon flux density (photons/sec·cm²)
A = junction area (cm²)
$H = \phi A$ = light intensity (photons/sec)



common application → optical receivers

- ## Light emitting diodes (LEDs)

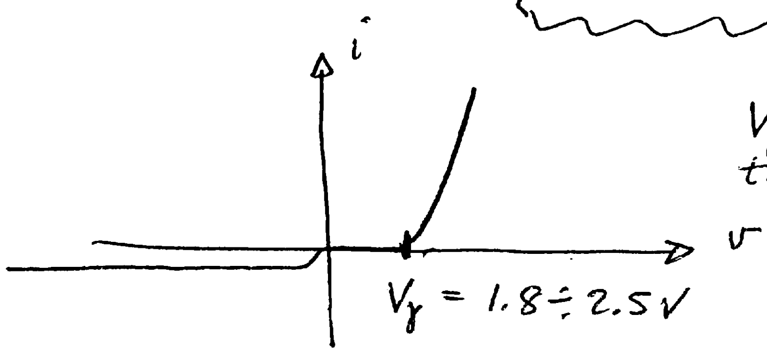


The LED performs the inverse of the function of the photodiodes
It converts a forward current (electric energy) into light.
The light emitted by a LED is proportional to the number of recombinations that take place and therefore is proportional to the current in the diode.

↓

When an electron fall from the conduction band into a hole give up energy in the form of light !!

GaAs emits light waves at a wavelength near the infrared band. To produce light in the visible range, GaP (gallium-phosphide) must be mixed with the GaAs. → ERROR ON TEXTBOOK P. 199



V_γ is equivalent to say the color of the light

LED are used in forward-bias!

LASER = LED designed so as to produce coherent light (very narrow bandwidth)

OPTOISOLATOR =